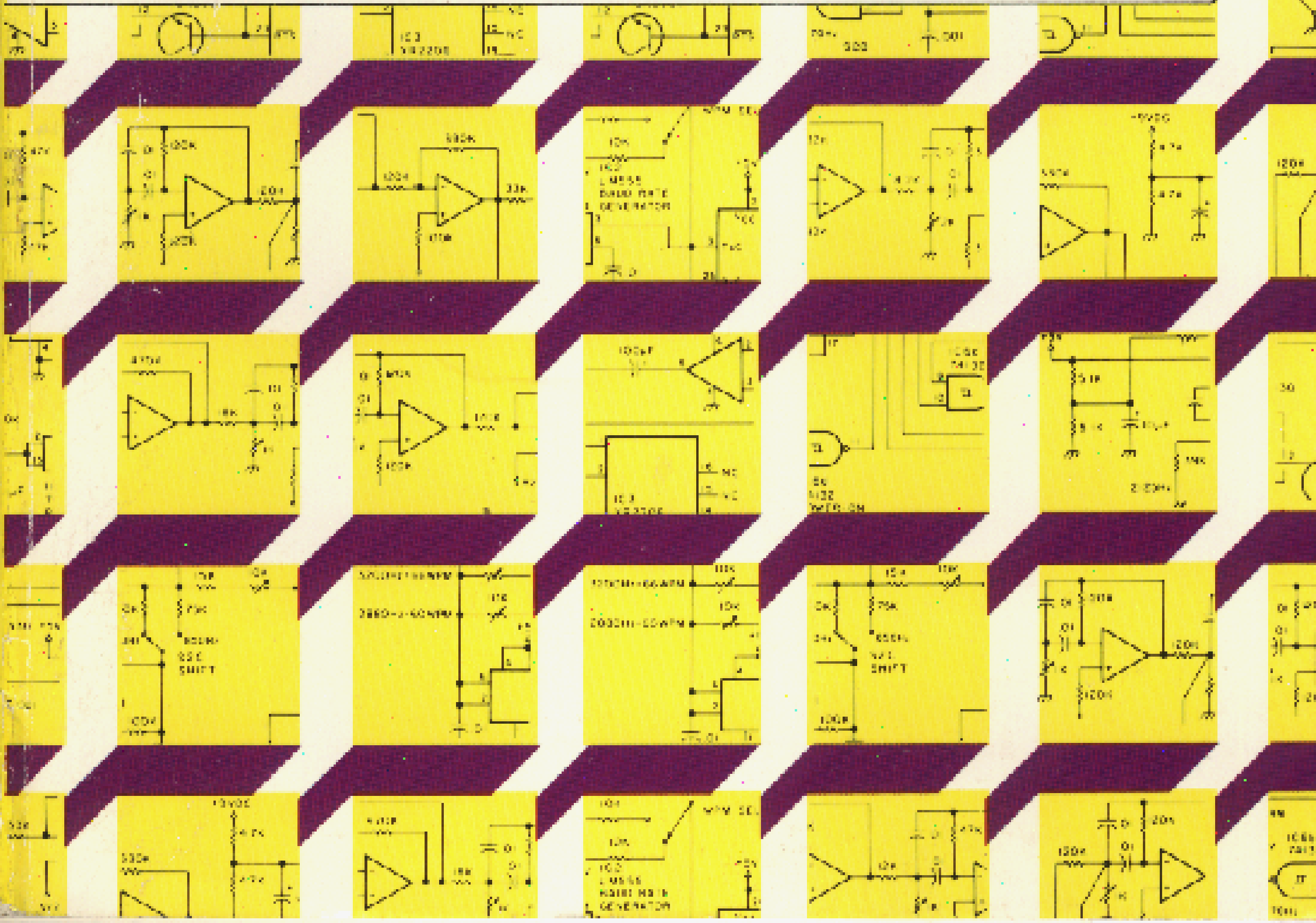
1938

ENCYCLOPEDIA OF ELECTRONIC CIRCUITS

VOLUME 1

RUDOLF F. GRAF



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Introduction

This volume of timely and practical circuits highlights the creative work of many people. Featured here are many circuits that appeared only briefly in some of our finer periodicals or limited-circulation publications. Also included are other useful and unique circuits from more readily available sources.

The source for each circuit is given in the sources section at the back of the book. The bold figure number that appears inside the box of each circuit is the key to the source. For example, the High Stability Voltage Reference circuit shown below is Fig. 93-10. If you turn to the Sources section and look for Fig. 93-10 you will find that Precision Monolithics supplied this circuit from p. 6-142 of their Full Line Catalog.

HIGH STABILITY VOLTAGE REFERENCE

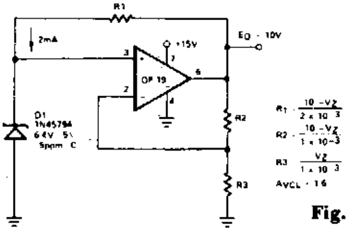


Fig. 93-10

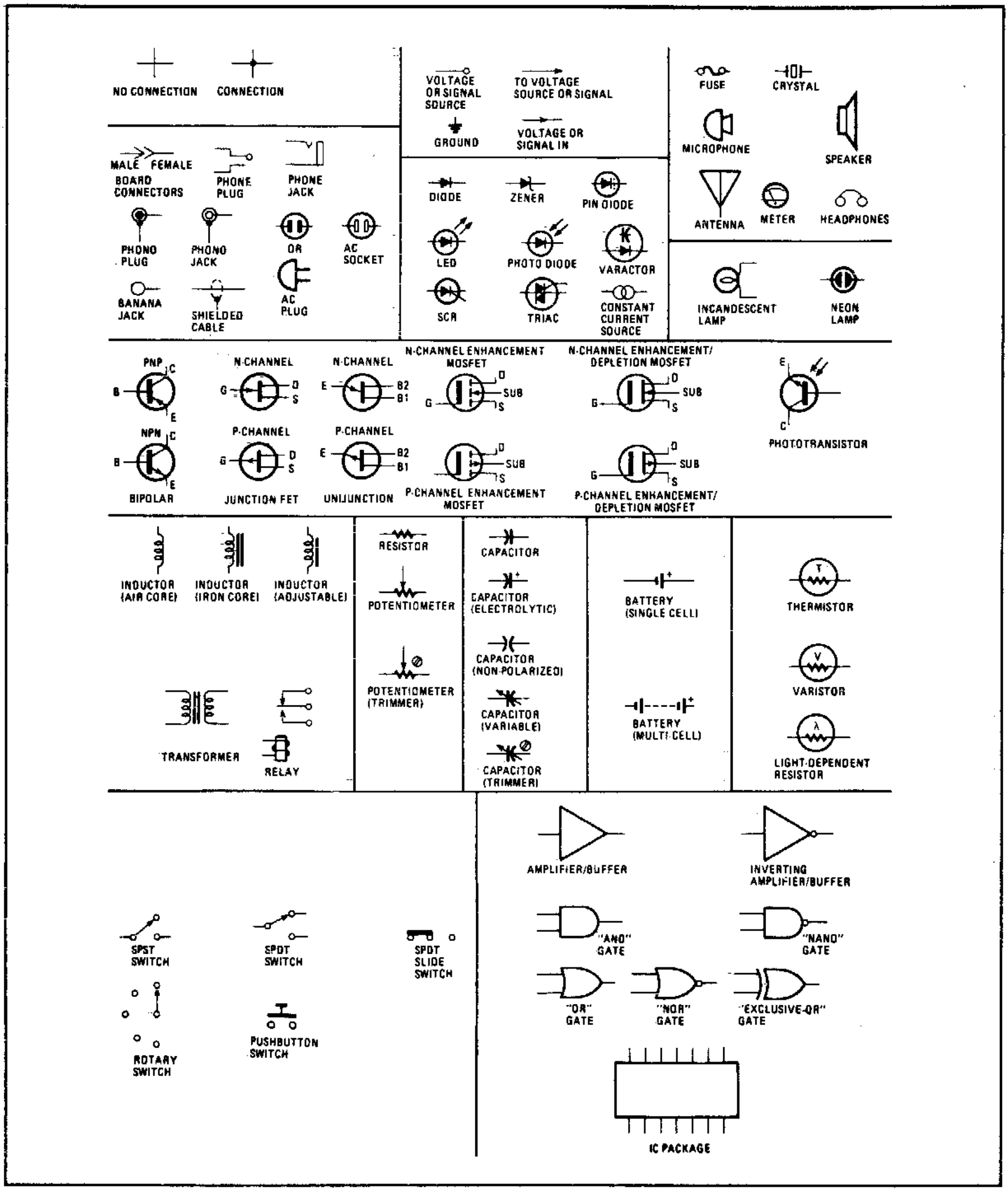
Fig. 93-9: Reprinted with the permission of National Semiconductor Corp. Transistor Databook, 1982, p. 11-25.
Fig. 93-10: Precision Monolithics Incorporated, 1981 Full Line Catalog, p. 6-142.
Fig. 93-11: Precision Monolithics Incorporated, 1981 Full Line Catalog, p. 10-18.

Many circuits are accompanied by a brief explanatory text. Those that do not have text can be readily understood from similar circuits in that chapter, or else they may be too complex to be explained briefly. The sparseness of text is deliberate so as to allow for more circuits which, after all, is what this book is all about.

The Index and Contents will be a time saver for the reader who knows exactly what he is looking for. The first page of each chapter lists the circuits in the order that they appear. The browser will surely discover many ideas and circuits that may well turn out to be most rewarding and great fun to put together.

The Common Schematic Symbols chart will help you identify circuit components.

Common Schematic Symbols


NO CONNECTION
CONNECTION
VOLTAGE OR SIGNAL SOURCE
TO VOLTAGE SOURCE OR SIGNAL
GROUND
VOLTAGE OR SIGNAL IN
FUSE
CRYSTAL
MICROPHONE
SPEAKER
MALE FEMALE
BOARD CONNECTORS
PHONE PLUG
PHONE JACK
PHONO PLUG
PHONO JACK
OR
AC SOCKET
AC PLUG
BANANA JACK
SHIELDED CABLE
DIODE
ZENER
PIN DIODE
LED
PHOTO DIODE
VARACTOR
SCR
TRIAC
CONSTANT CURRENT SOURCE
ANTENNA
METER
HEADPHONES
INCANDESCENT LAMP
NEON LAMP
PNP
C
B
E
NPN
BIPOLAR
N-CHANNEL
G
D
S
P-CHANNEL
JUNCTION FET
N-CHANNEL
E
B2
B1
P-CHANNEL
UNIJUNCTION
N-CHANNEL ENHANCEMENT MOSFET
SUB
P-CHANNEL ENHANCEMENT MOSFET
N-CHANNEL ENHANCEMENT/ DEPLETION MOSFET
P-CHANNEL ENHANCEMENT/ DEPLETION MOSFET
PHOTOTRANSISTOR
INDUCTOR (AIR CORE)
INDUCTOR (IRON CORE)
INDUCTOR (ADJUSTABLE)
TRANSFORMER
RELAY
RESISTOR
POTENTIOMETER
POTENTIOMETER (TRIMMER)
CAPACITOR
CAPACITOR (ELECTROLYTIC)
CAPACITOR (NON-POLARIZED)
CAPACITOR (VARIABLE)
CAPACITOR (TRIMMER)
BATTERY (SINGLE CELL)
BATTERY (MULTI-CELL)
THERMISTOR
VARISTOR
LIGHT-DEPENDENT RESISTOR
SPST SWITCH
SPDT SWITCH
SPDT SLIDE SWITCH
ROTARY SWITCH
PUSHBUTTON SWITCH
AMPLIFIER/BUFFER
INVERTING AMPLIFIER/BUFFER
"AND" GATE
"NAND" GATE
"OR" GATE
"NOR" GATE
"EXCLUSIVE-OR" GATE
IC PACKAGE

1

Alarms

The sources of the following circuits are contained in the Sources section beginning on page 730. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

Computalarm
Automotive Burglar Alarm
Security Alarm
Vehicle Security System
Home Security Monitor System
Antitheft Device
Auto Burglar Alarm
Tamper-Proof Burglar Alarm
Latching Burglar Alarm
Motion-Activated Motorcycle or Car Alarm
Boat Alarm
Blown Fuse Alarm
Auto Burglar Alarm
Continuous-Tone 2 kHz Buzzer with Bridge Drive, Gated on by a Logic 0
Pulsed-Tone Alarm, Gated by a High Input, with Direct-Drive Output
Piezoelectric Alarm
Gated 2 kHz Buzzer
Burglar Alarm
Latching Burglar Alarm
Sun-Powered Alarm
Freezer Meltdown Alarm

COMPUTALARM

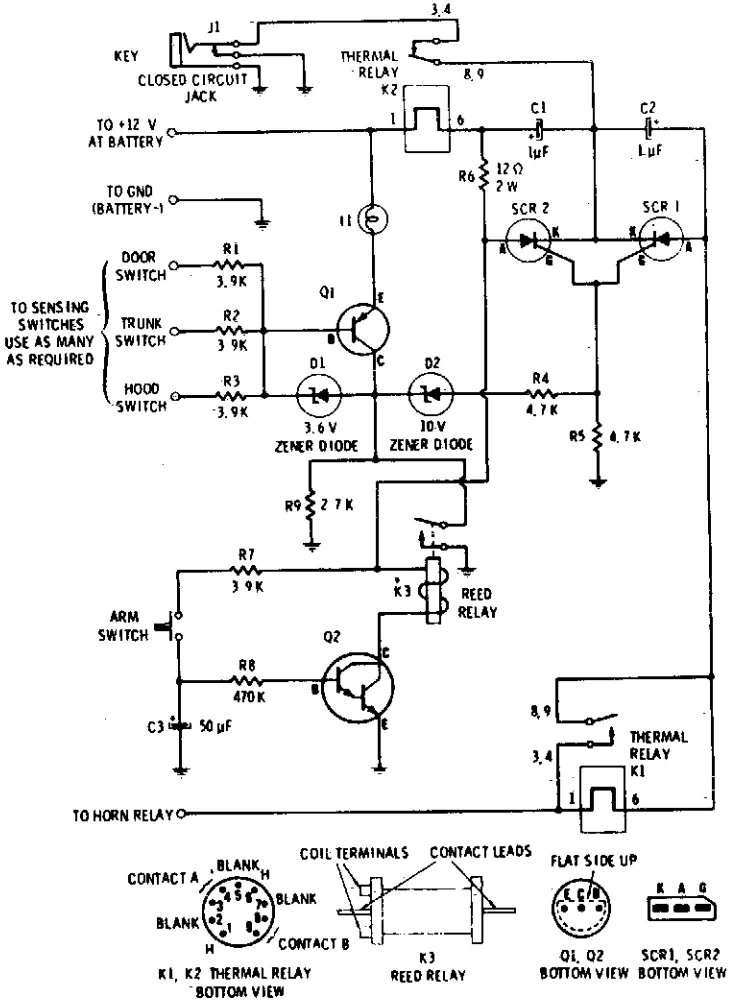

J1
KEY
CLOSED CIRCUIT JACK
3,4
THERMAL RELAY K2
8,9
TO +12 V AT BATTERY
1
6
C1 1µF
C2 1µF
R6 12 Ω 2 W
TO GND (BATTERY -)
I1
SCR 2
SCR 1
A
K
G
DOOR SWITCH
R1 3.9K
TO SENSING SWITCHES USE AS MANY AS REQUIRED
TRUNK SWITCH
R2 3.9K
Q1
B
E
C
HOOD SWITCH
R3 3.9K
D1 3.6 V ZENER DIODE
D2 10 V ZENER DIODE
R4 4.7K
R5 4.7K
R9 2.7K
R7 3.9K
K3 REED RELAY
ARM SWITCH
Q2
R8 470K
C3 50 µF
8,9
3,4
THERMAL RELAY K1
TO HORN RELAY
CONTACT A
BLANK
H
BLANK
BLANK
CONTACT B
K1, K2 THERMAL RELAY BOTTOM VIEW
COIL TERMINALS
CONTACT LEADS
K3 REED RELAY
FLAT SIDE UP
Q1, Q2 BOTTOM VIEW
K A G
SCR1, SCR2 BOTTOM VIEW

Fig. 1-1

Circuit Notes

The circuit has a built-in, self-arming feature. The driver turns off the ignition, presses the arm button on the Computalarm, and leaves the car. Within 20 seconds, the alarm arms itself—all automatically! The circuit will then detect the opening of any monitored door, the trunk lid, or the hood on the car. Once activated, the circuit remains dormant for 10 seconds. When the 10-second time delay has run out, the circuit will close the car's horn relay and sound the horn in periodic blasts (approximately 1 to 2 seconds apart) for a period of one minute. Then the Computalarm automatically shuts itself off (to save your battery) and rearms. If a door, the trunk lid, or the hood remains ajar, the alarm circuit retriggers and another period of horn blasts occurs. The Computalarm has a "key" switch by which the driver can disarm the alarm circuit within a 10-second period after he enters the door. The key switch consists of a closed circuit jack, J1, and a mating miniature plug.

AUTOMOTIVE BURGLAR ALARM

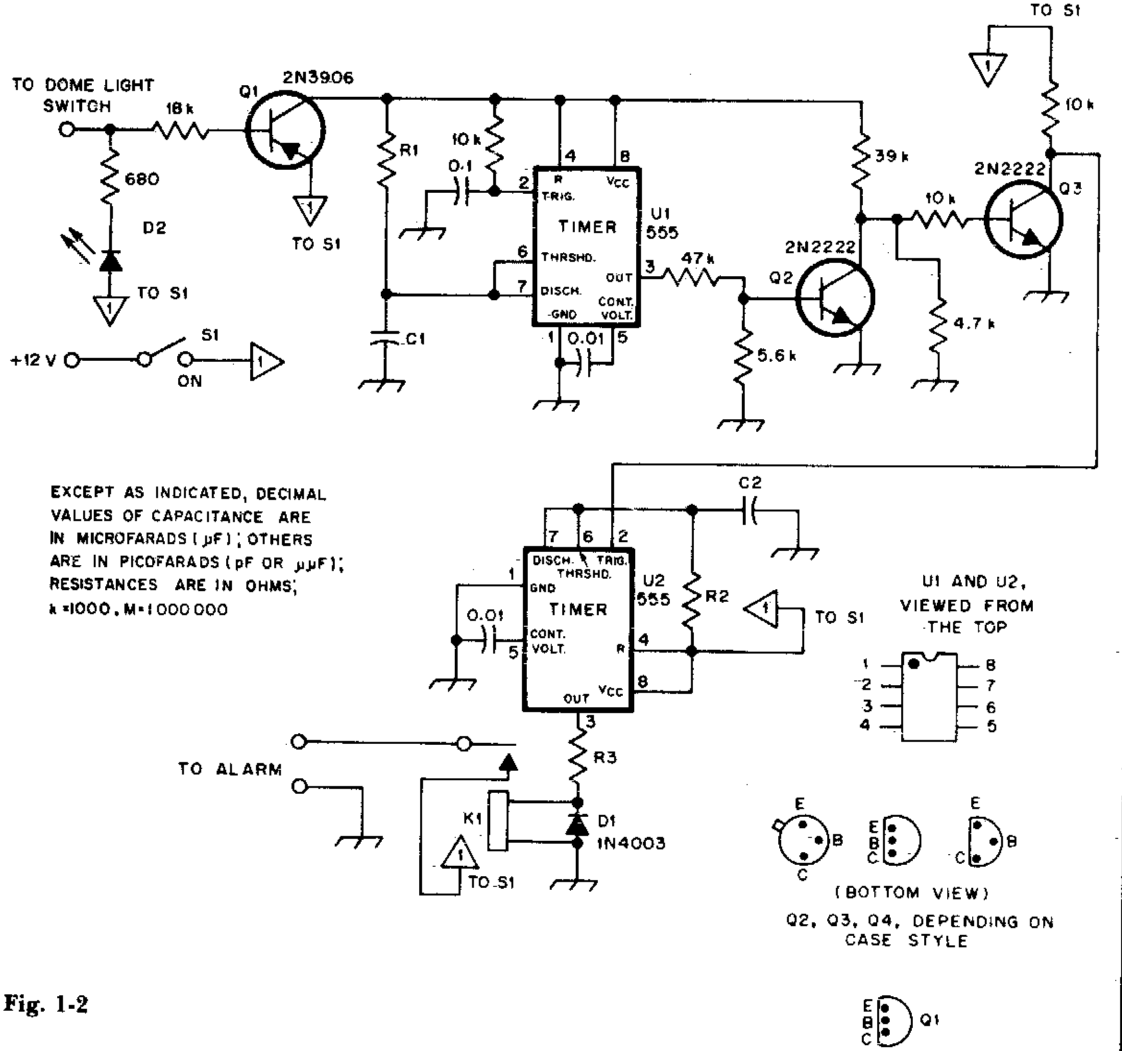

TO S1
TO DOME LIGHT SWITCH
18 k
Q1
2N3906
680
D2
TO S1
TO S1
+12 V
S1
ON
R1
10 k
0.1
C1
4
8
2
R
VCC
TRIG.
TIMER
U1
555
6
THRSHD.
7
DISCH.
OUT
3
GND
CONT. VOLT.
1
0.01
5
47 k
5.6 k
39 k
Q2
2N2222
10 k
4.7 k
10 k
2N2222
Q3
EXCEPT AS INDICATED, DECIMAL VALUES OF CAPACITANCE ARE IN MICROFARADS (µF); OTHERS ARE IN PICOFARADS (pF OR µµF); RESISTANCES ARE IN OHMS; k =1000, M=1 000 000
C2
7
6
2
DISCH.
TRIG.
THRSHD.
GND
U2
555
TIMER
R2
TO S1
0.01
CONT. VOLT.
5
R
4
8
OUT
VCC
3
R3
TO ALARM
K1
D1
1N4003
TO S1
U1 AND U2, VIEWED FROM THE TOP
1
2
3
4
8
7
6
5
E
B
C
E
B
C
E
C
B
(BOTTOM VIEW)
Q2, Q3, Q4, DEPENDING ON CASE STYLE
E
B
C
Q1

Fig. 1-2

Circuit Notes

Alarm triggers on after a 13 second delay and stays on for 1-1½ minutes. Then it resets automatically. It can also be turned off and reset by opening and reclosing S1.

SECURITY ALARM

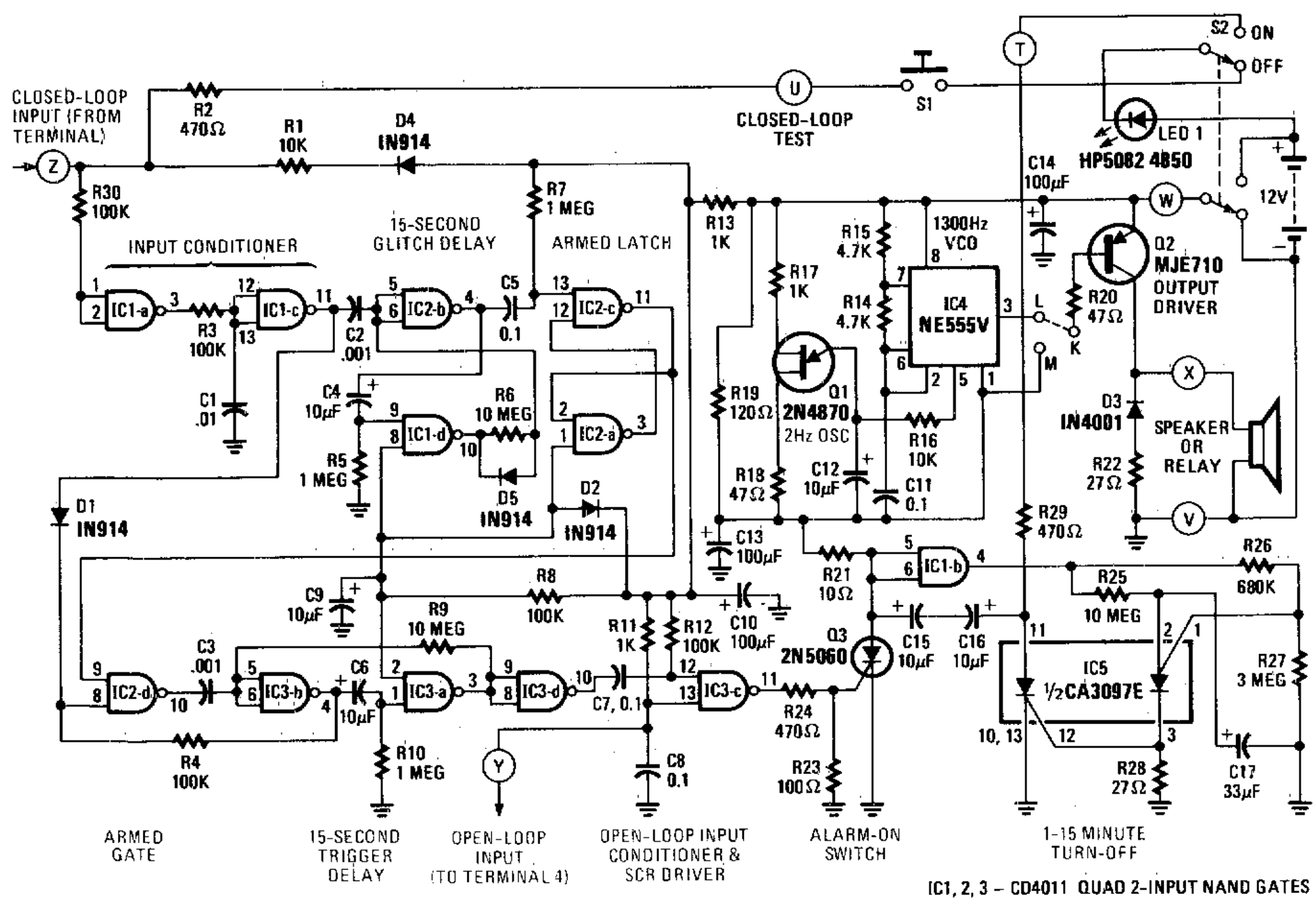
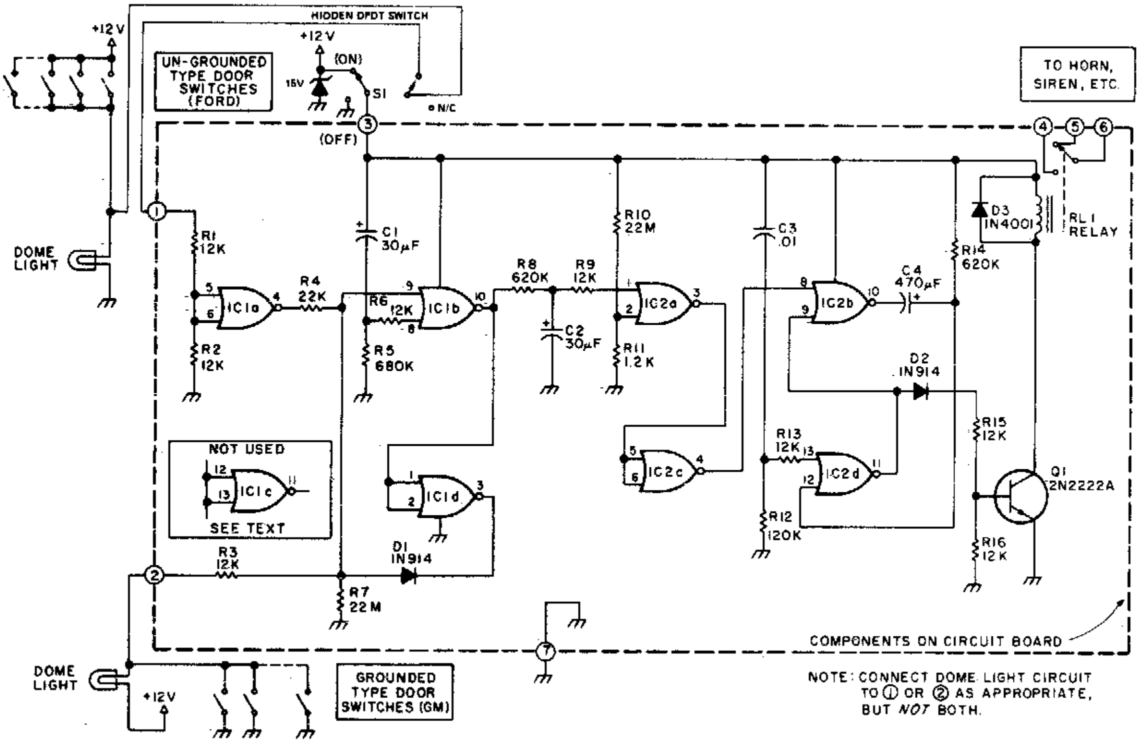

CLOSED-LOOP INPUT (FROM TERMINAL)
Z
R30 100K
R2 470Ω
R1 10K
D4 IN914
U
CLOSED-LOOP TEST
S1
T
S2 ON OFF
LED 1 HP5082 4850
12V
W
C14 100μF
R13 1K
R7 1 MEG
INPUT CONDITIONER
15-SECOND GLITCH DELAY
ARMED LATCH
IC1-a
IC1-c
R3 100K
C1 .01
C2 .001
IC2-b
C5 0.1
IC2-c
IC2-a
C4 10μF
IC1-d
R5 1 MEG
R6 10 MEG
D5 IN914
D2 IN914
D1 IN914
R15 4.7K
R14 4.7K
R17 1K
1300Hz VCO
IC4 NE555V
R19 120Ω
Q1 2N4870
2Hz OSC
R16 10K
R18 47Ω
C12 10μF
C11 0.1
C13 100μF
L
K
M
R20 47Ω
Q2 MJE710 OUTPUT DRIVER
X
D3 IN4001
SPEAKER OR RELAY
R22 27Ω
V
R29 470Ω
R21 10Ω
IC1-b
R26 680K
R25 10 MEG
C9 10μF
R8 100K
C10 100μF
R11 1K
R12 100K
Q3 2N5060
C15 10μF
C16 10μF
IC5 ½CA3097E
R27 3 MEG
IC2-d
C3 .001
IC3-b
C6 10μF
IC3-a
R9 10 MEG
IC3-d
C7, 0.1
IC3-c
R24 470Ω
R28 27Ω
C17 33μF
R4 100K
R10 1 MEG
Y
C8 0.1
R23 100Ω
ARMED GATE
15-SECOND TRIGGER DELAY
OPEN-LOOP INPUT (TO TERMINAL 4)
OPEN-LOOP INPUT CONDITIONER & SCR DRIVER
ALARM-ON SWITCH
1-15 MINUTE TURN-OFF
IC1, 2, 3 – CD4011 QUAD 2-INPUT NAND GATES

Fig. 1-3

Circuit Notes

This alarm features open- and closed-loop detector and automatic alarm shutoff. Offers 15 second exit/entrance delay. Alarm on time can be adjusted from 1 to 15 minutes.

VEHICLE SECURITY SYSTEM



+12V
HIDDEN DPDT SWITCH
+12V
(ON)
15V
S1
N/C
(OFF)
UN-GROUNDED TYPE DOOR SWITCHES (FORD)
TO HORN, SIREN, ETC.
DOME LIGHT
R1 12K
R2 12K
R4 22K
IC1a
IC1b
R6 12K
R5 680K
C1 30µF
R8 620K
R9 12K
R10 22M
C2 30µF
R11 1.2K
IC2a
C3 .01
IC2b
C4 470µF
R14 620K
D3 1N4001
RL1 RELAY
D2 1N914
R13 12K
IC2c
IC2d
R12 120K
R15 12K
Q1 2N2222A
R16 12K
NOT USED
IC1c
SEE TEXT
IC1d
R3 12K
D1 1N914
R7 22M
COMPONENTS ON CIRCUIT BOARD
DOME LIGHT
+12V
GROUNDED TYPE DOOR SWITCHES (GM)
NOTE: CONNECT DOME LIGHT CIRCUIT TO ① OR ② AS APPROPRIATE, BUT NOT BOTH.

Fig. 1-4

Circuit Notes

This alarm gives a 15-20 second exit and entrance delay. After being triggered, the alarm sounds for five minutes and then shuts off. Once triggered, the sequence is automatic and is not affected by subsequent opening or closing of doors.

HOME SECURITY MONITOR SYSTEM

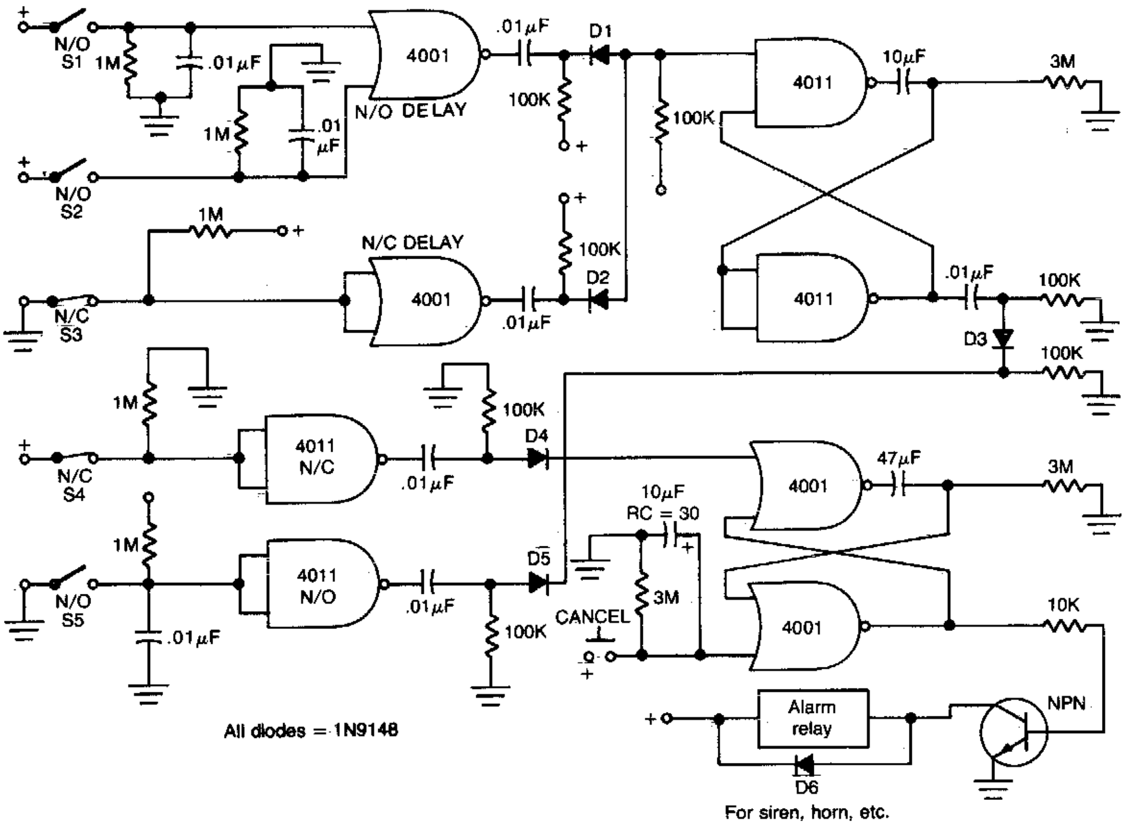

+
N/O
S1
1M
.01μF
4001
N/O DELAY
.01μF
D1
100K
+
4011
10μF
3M
100K
+
1M
.01
μF
N/O
S2
1M
+
N/C DELAY
+
100K
D2
4001
.01μF
4011
.01μF
100K
N/C
S3
D3
100K
1M
100K
D4
+
4011
N/C
.01μF
N/C
S4
4001
47μF
3M
10μF
RC = 30
1M
D5
+
4011
N/O
.01μF
3M
N/O
S5
CANCEL
100K
4001
10K
.01μF
+
All diodes = 1N9148
+
Alarm
relay
NPN
D6
For siren, horn, etc.

Fig. 1-5

Circuit Notes

This circuit provides normally open (NO) and normally closed (NC) contacts S1, S2, and S3 to turn on the alarm after a 30 second delay. S4 and S5 operate instantly. The CANCEL switch resets the alarm.

ANTITHEFT DEVICE

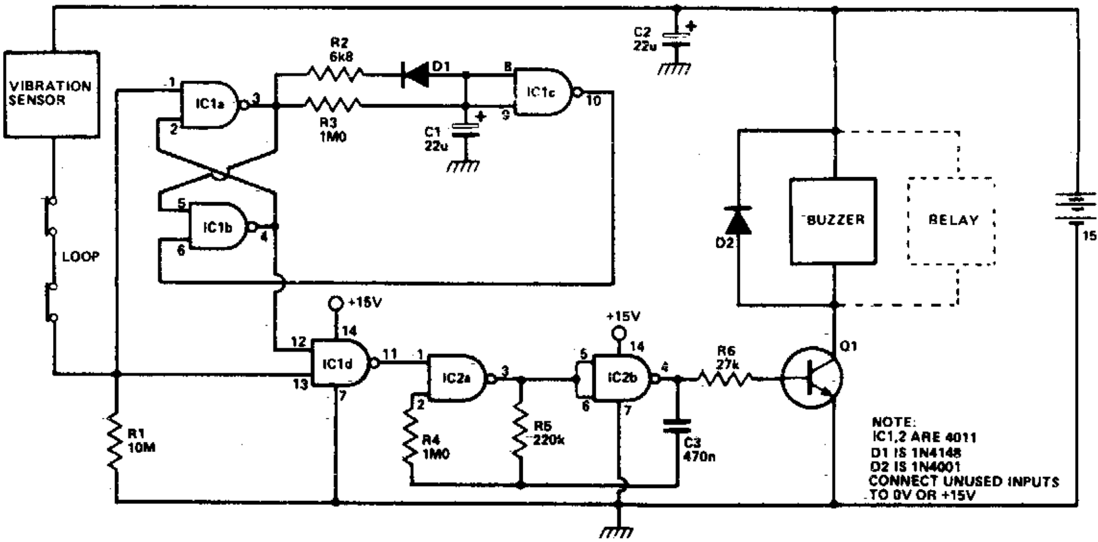

VIBRATION SENSOR
LOOP
R2 6k8
D1
R3 1M0
C1 22u
C2 22u
IC1a
IC1b
IC1c
IC1d
IC2a
IC2b
+15V
R1 10M
R4 1M0
R5 220k
R6 27k
C3 470n
Q1
D2
BUZZER
RELAY
15V
NOTE:
IC1,2 ARE 4011
D1 IS 1N4148
D2 IS 1N4001
CONNECT UNUSED INPUTS
TO 0V OR +15V

Fig. 1-6

Circuit Notes

Any momentary break in the protective loop or tripping of the normally closed vibration sensor, causes alarm to sound for 20 seconds. If the circuit is open all the time, the alarm will sound continuously.

AUTO BURGLAR ALARM

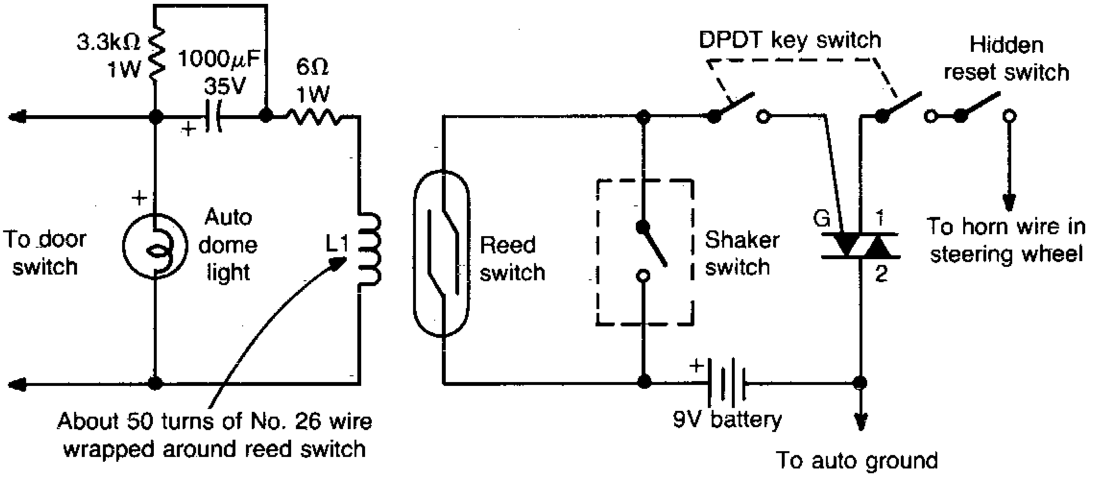

3.3kΩ 1W
1000μF 35V
6Ω 1W
To door switch
Auto dome light
L1
About 50 turns of No. 26 wire wrapped around reed switch
Reed switch
Shaker switch
DPDT key switch
Hidden reset switch
G
1
2
To horn wire in steering wheel
9V battery
To auto ground

Fig. 1-7

Circuit Notes

Dome light current through L1 closes reed switch and sounds alarm. Shaker switch also activates alarm.

TAMPER-PROOF BURGLAR ALARM

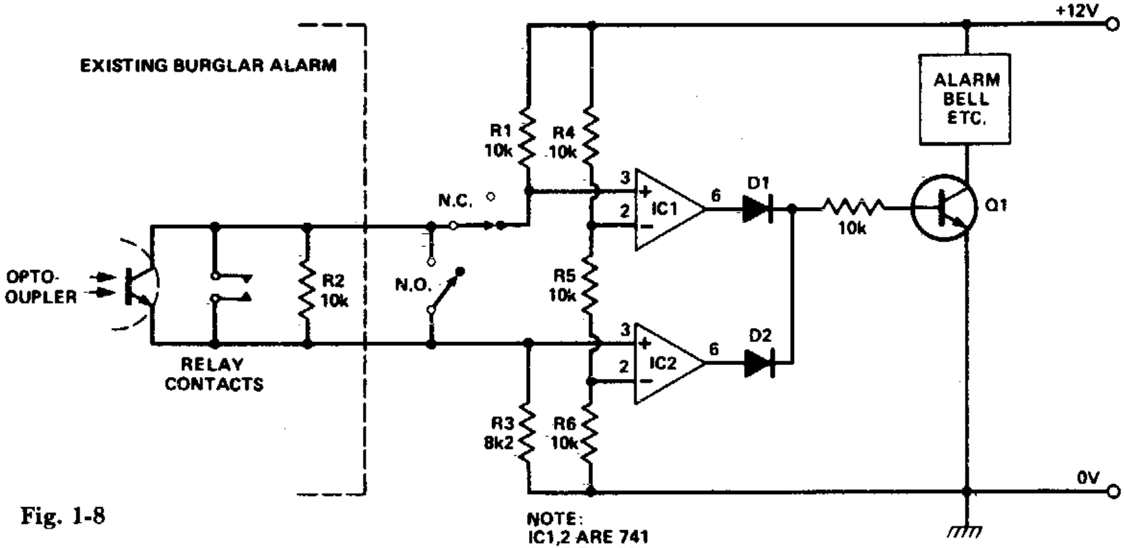

+12V
EXISTING BURGLAR ALARM
ALARM BELL ETC.
R1 10k
R4 10k
N.C.
3
2
IC1
6
D1
10k
Q1
OPTO-OUPLER
R2 10k
N.O.
R5 10k
3
2
IC2
6
D2
RELAY CONTACTS
R3 8k2
R6 10k
0V

Fig. 1-8

NOTE:
IC1,2 ARE 741
Q1 IS TIP122
D1,2 ARE GENERAL PURPOSE DIODES

Circuit Notes

If R2 is opened or shorted, the alarm sounds.

LATCHING BURGLAR ALARM

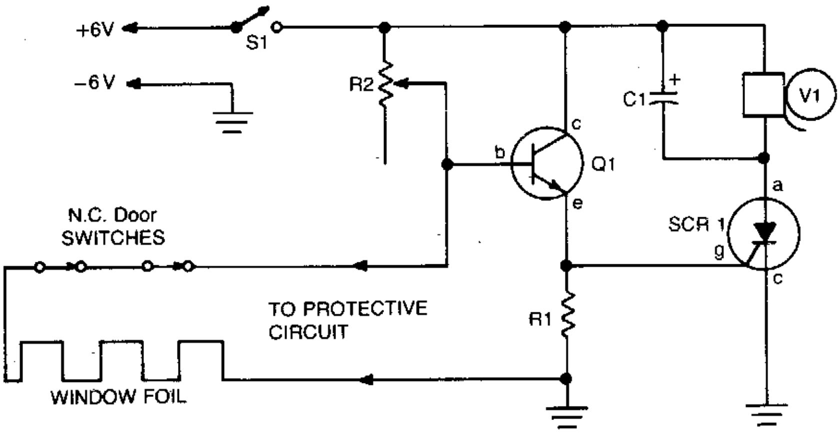

+6V
S1
−6V
R2
C1
+
V1
c
b
Q1
e
a
N.C. Door SWITCHES
SCR 1
g
c
TO PROTECTIVE CIRCUIT
R1
WINDOW FOIL

Fig. 1-9

Circuit Notes

When the protective circuit is interrupted (opened), the alarm sounds. To set the circuit, adjust R2 (with protective circuit open) for 1 V across R1.

MOTION-ACTIVATED MOTORCYCLE OR CAR ALARM

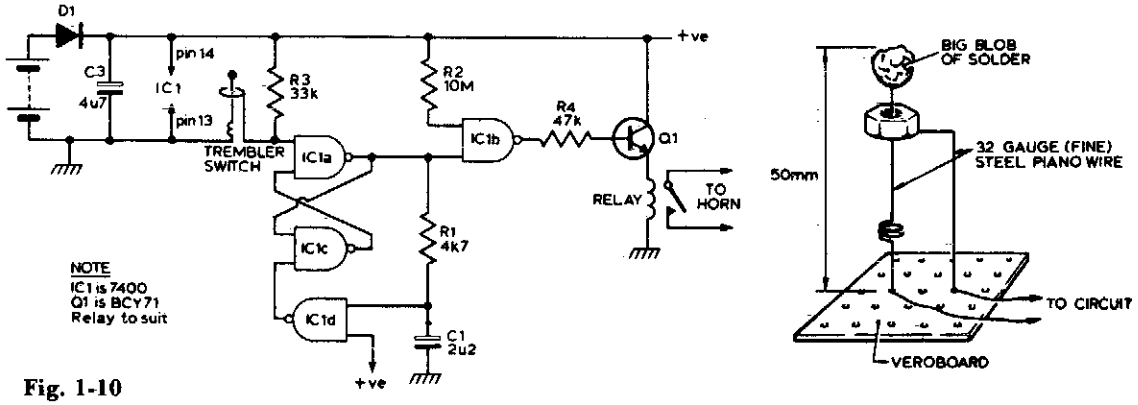
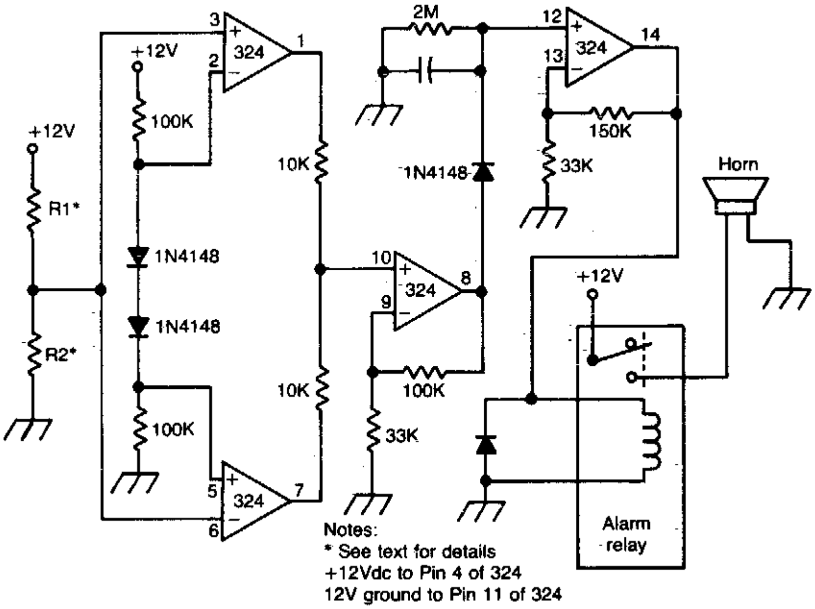

D1
C3
4u7
pin 14
IC1
pin 13
TREMBLER SWITCH
R3
33k
R2
10M
IC1a
IC1b
IC1c
IC1d
R4
47k
Q1
+ve
RELAY
TO HORN
R1
4k7
C1
2u2
+ve
NOTE
IC1 is 7400
Q1 is BCY71
Relay to suit
BIG BLOB OF SOLDER
32 GAUGE (FINE) STEEL PIANO WIRE
50mm
TO CIRCUIT
VEROBOARD

Fig. 1-10

Circuit Notes

Trembler (motion activated) switch sounds the alarm for 5 seconds. Then it goes off. Circuit is timed out for 10 seconds to allow the trembler switch to settle.

BOAT ALARM


+12V
R1*
R2*
+12V
100K
1N4148
1N4148
100K
3
2
324
1
10K
10K
5
6
324
7
2M
1N4148
10
9
324
8
100K
33K
12
13
324
14
150K
33K
Horn
+12V
Alarm relay

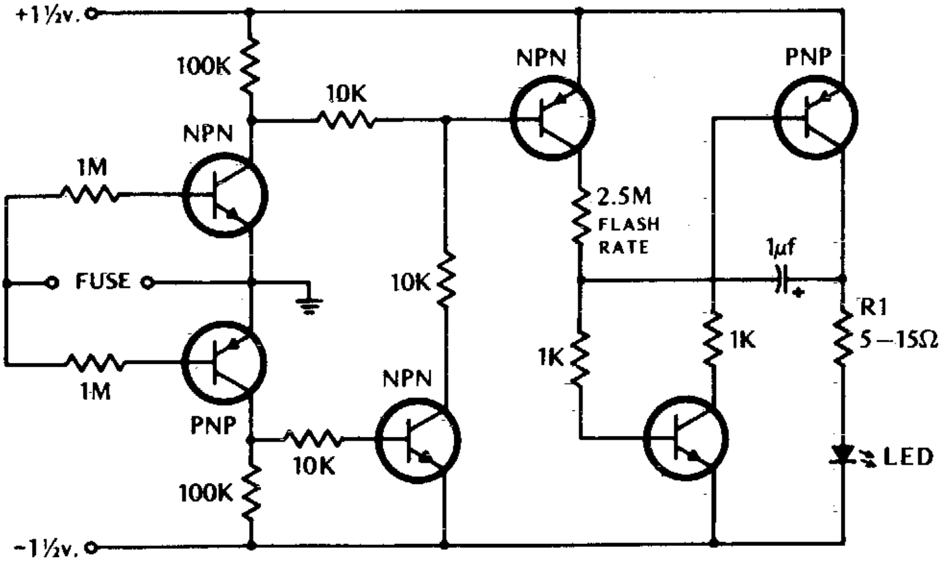
Notes:
* See text for details
+12Vdc to Pin 4 of 324
12V ground to Pin 11 of 324

Fig. 1-11

Circuit Notes

Removing R1 or R2 from the circuit (i.e., the potential thief breaks a hidden wire that connects R1 to +12 V and R2 to ground) activates the alarm for about five minutes.

BLOWN-FUSE ALARM



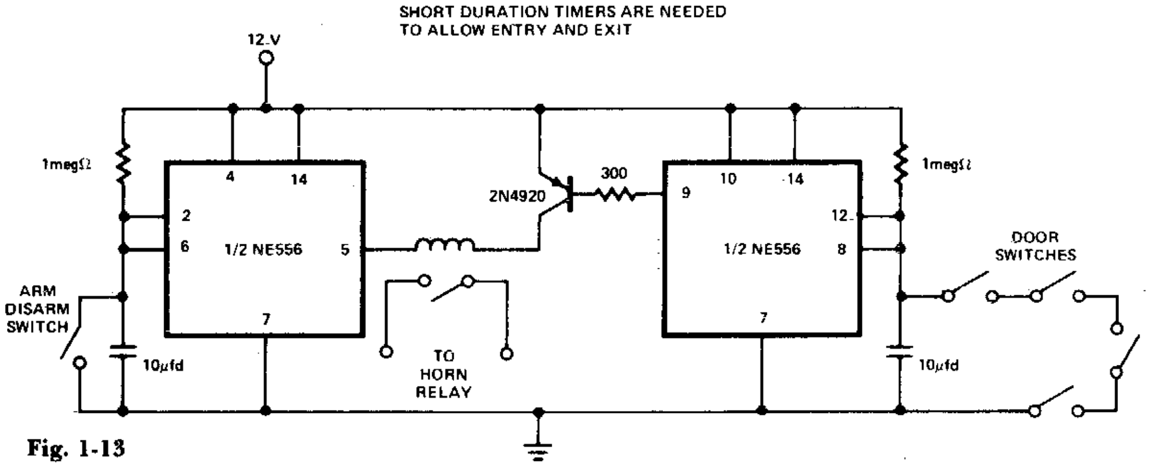
+1½v.
100K
10K
NPN
1M
FUSE
NPN
PNP
10K
2.5M
FLASH
RATE
1µf
R1
5–15Ω
1K
1K
1M
NPN
PNP
10K
100K
LED
−1½v.

Fig. 1-12

Circuit Notes

If the fuse blows, the LED indicator starts to blink.

AUTO BURGLAR ALARM



SHORT DURATION TIMERS ARE NEEDED
TO ALLOW ENTRY AND EXIT
12 V
1megΩ
4
14
2
6
1/2 NE556
5
7
2N4920
300
9
10
14
12
8
1/2 NE556
7
1megΩ
DOOR
SWITCHES
ARM
DISARM
SWITCH
10µfd
TO
HORN
RELAY
10µfd

Fig. 1-13

CONTINUOUS-TONE 2 kHz BUZZER WITH BRIDGE DRIVE, GATED ON BY A LOGIC 0

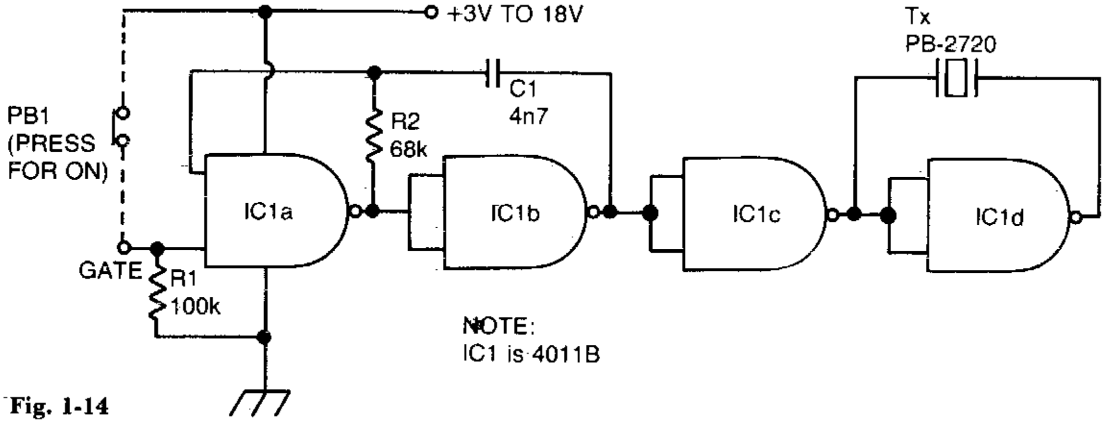

+3V TO 18V
Tx
PB-2720
C1
4n7
R2
68k
PB1
(PRESS
FOR ON)
IC1a
IC1b
IC1c
IC1d
GATE
R1
100k
NOTE:
IC1 is 4011B

Fig. 1-14

PULSED-TONE ALARM, GATED BY A HIGH INPUT, WITH DIRECT-DRIVE OUTPUT

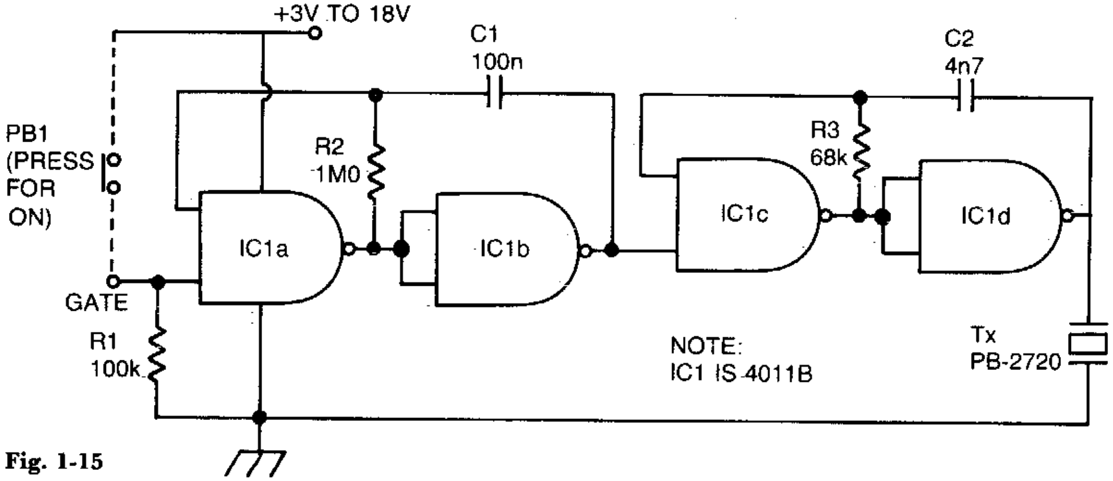

+3V TO 18V
C1
100n
C2
4n7
PB1
(PRESS
FOR
ON)
R2
1M0
R3
68k
IC1a
IC1b
IC1c
IC1d
GATE
R1
100k
NOTE:
IC1 IS 4011B
Tx
PB-2720

Fig. 1-15

PIEZOELECTRIC ALARM

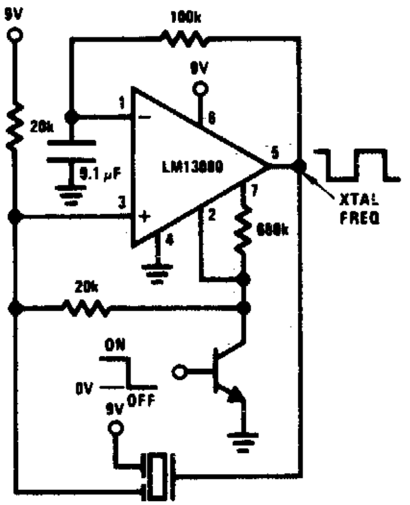

9V
100k
9V
1
6
20k
0.1 µF
LM13080
5
3
7
XTAL FREQ
2
680k
4
20k
ON
0V
OFF
9V

Fig. 1-16

BURGLAR ALARM

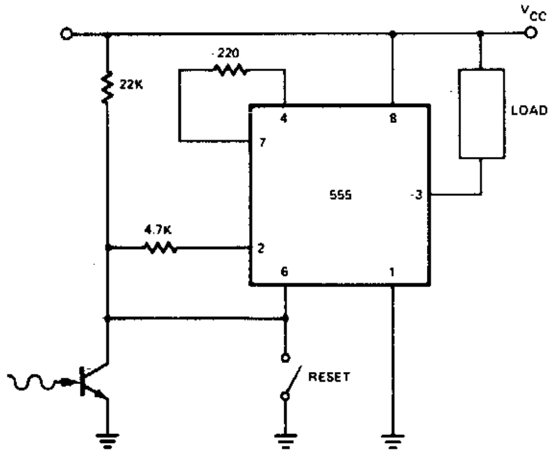
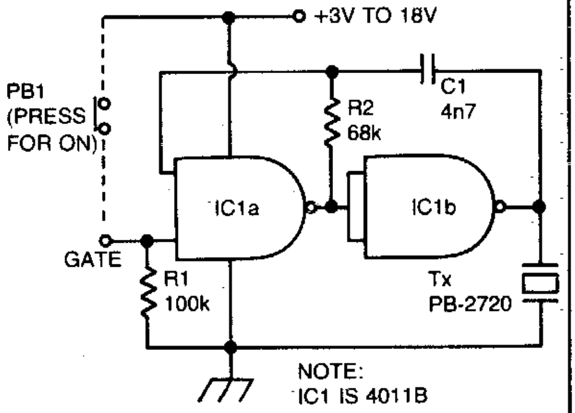

VCC
220
22K
4
8
LOAD
7
555
3
4.7K
2
6
1
RESET

Fig. 1-18

GATED 2 kHz BUZZER


+3V TO 18V
C1
4n7
R2
68k
PB1
(PRESS
FOR ON)
IC1a
IC1b
GATE
R1
100k
Tx
PB-2720

NOTE:
IC1 IS 4011B

Fig. 1-17

LATCHING BURGLAR ALARM

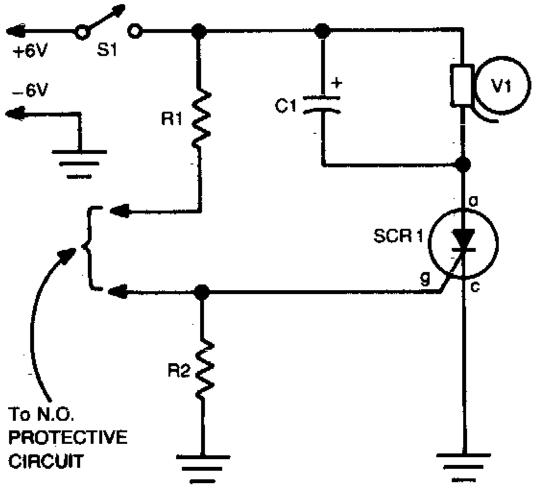

+6V
S1
−6V
R1
C1
+
V1
a
SCR1
g
c
R2
To N.O.
PROTECTIVE
CIRCUIT

Fig. 1-19

Circuit Notes

Closing the protective circuit (i.e., R1 to R2) applies positive voltage to the gate of SCR1 and sounds the alarm. It can only be turned off with S1.

SUN-POWERED ALARM

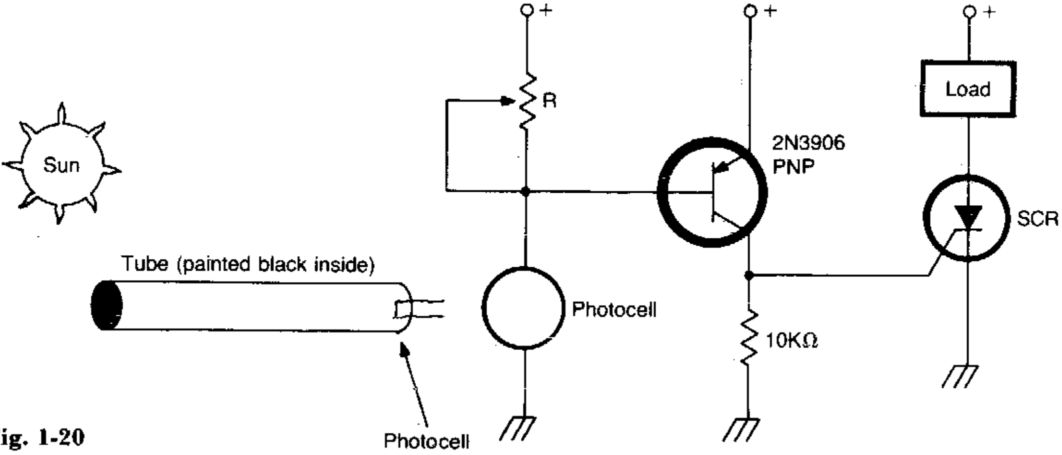

+
+
+
R
Load
Sun
2N3906
PNP
SCR
Tube (painted black inside)
Photocell
10KΩ
Photocell

Fig. 1-20

Circuit Notes

Circuit turns on when light (sunlight) strikes photocell. Potentiometer R sets light level at which the alarm sounds. Painted tube (black on inside) may be used on photocell to aim at the sun.

FREEZER MELTDOWN ALARM

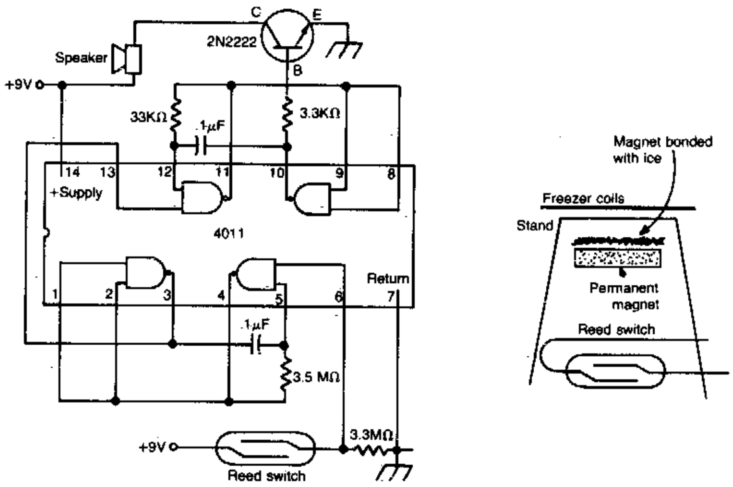

C
E
2N2222
B
Speaker
+9V
33KΩ
.1μF
3.3KΩ
14
13
12
11
10
9
8
+Supply
4011
1
2
3
4
5
6
Return
7
.1μF
3.5 MΩ
3.3MΩ
+9V
Reed switch
Magnet bonded with ice
Freezer coils
Stand
Permanent magnet
Reed switch

Fig. 1-21

Circuit Notes

The meltdown is a magnet held to a small stand by ice. A reed switch is below the magnet. When the ice melts, the magnet falls on the switch, closing it, and completing the alarm circuit.

2

Amateur Radio

The sources of the following circuits are contained in the Sources section beginning on page 730. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

Code Practice Oscillator Produces Automatic Dits and Dahs
Rf Power Meter
In-Line Wattmeter
CW Signal Processor
Two-Meter Preamplifier for Handitalkies
Repeater Beeper
Electronic Keyer
Code Practice Oscillator
Automatic Tape Recording
Self-Powered CW Monitor
Remote Rf Current Readout
Code Practice Oscillator
SWR Warning Indicator
Subaudible Tone Encoder
Audio Mixers
Rf Powered Sidetone Oscillator
Harmonic Generator
Automatic TTL Morse-Code Keyer
Remote Rf Current Readout

CODE-PRACTICE OSCILLATOR PRODUCES AUTOMATIC DITS AND DAHS

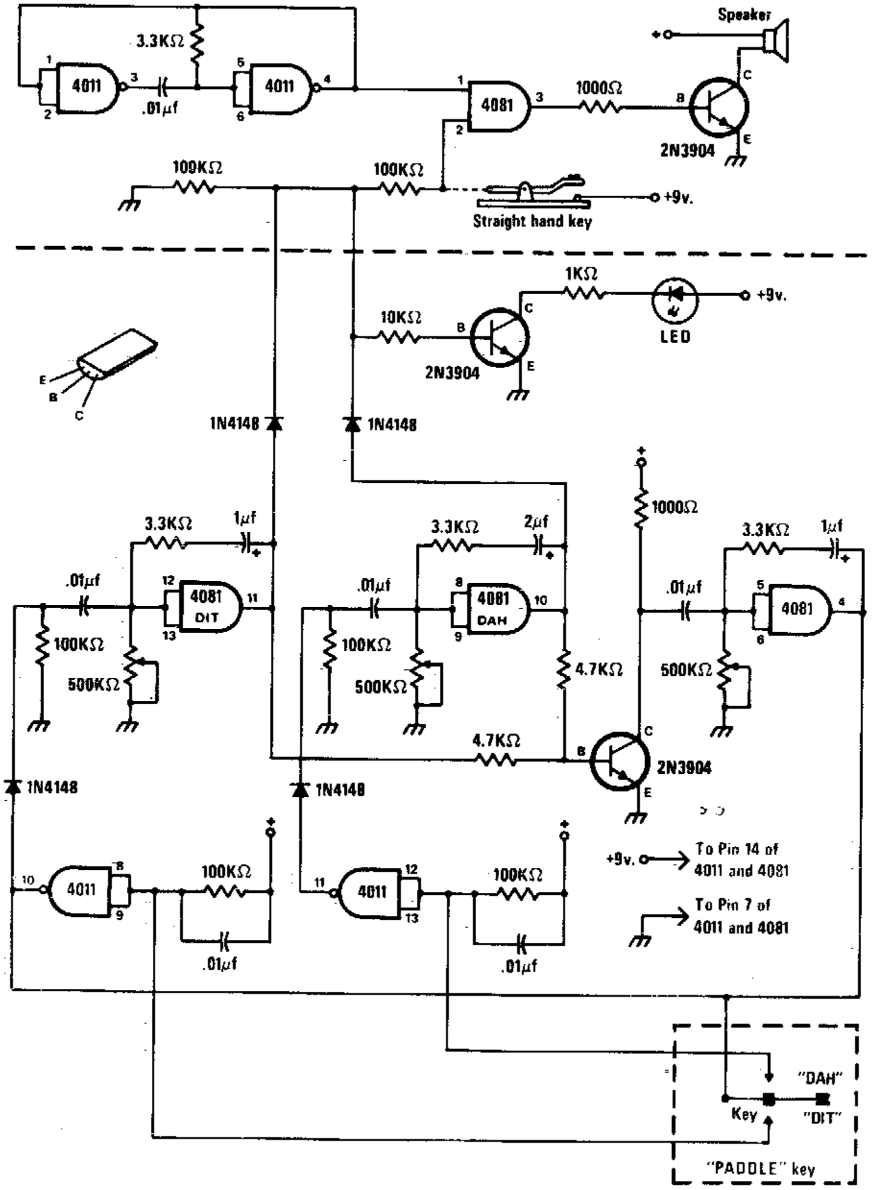

Speaker
+
3.3KΩ
4011
4011
.01μf
1
2
3
5
6
4
4081
1
2
3
1000Ω
B
C
E
2N3904
100KΩ
100KΩ
Straight hand key
+9v.
1KΩ
10KΩ
B
C
E
2N3904
LED
+9v.
E
B
C
1N4148
1N4148
3.3KΩ
1μf
.01μf
12
13
11
4081
DIT
100KΩ
500KΩ
3.3KΩ
2μf
.01μf
8
9
10
4081
DAH
100KΩ
500KΩ
+
1000Ω
3.3KΩ
1μf
.01μf
5
6
4
4081
500KΩ
4.7KΩ
4.7KΩ
B
C
E
2N3904
1N4148
1N4148
10
8
9
4011
100KΩ
.01μf
+
11
12
13
4011
100KΩ
.01μf
+
+9v.
To Pin 14 of 4011 and 4081
To Pin 7 of 4011 and 4081
"DAH"
Key
"DIT"
"PADDLE" key

Fig. 2-1

Circuit Notes

The circuit consists of a basic oscillator (above dashed line) and an automatic keyer (below dashed line). The unit can be used with a straight hand key or a paddle key for automatic operation.

RF POWER METER

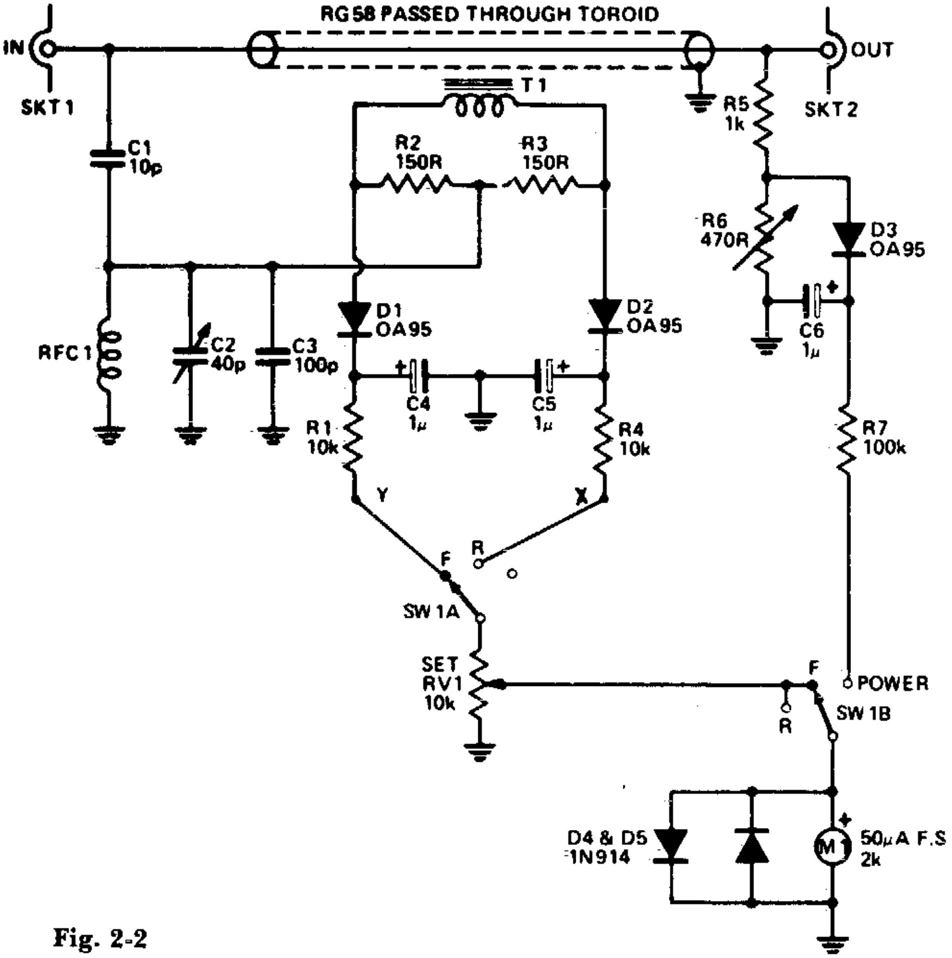
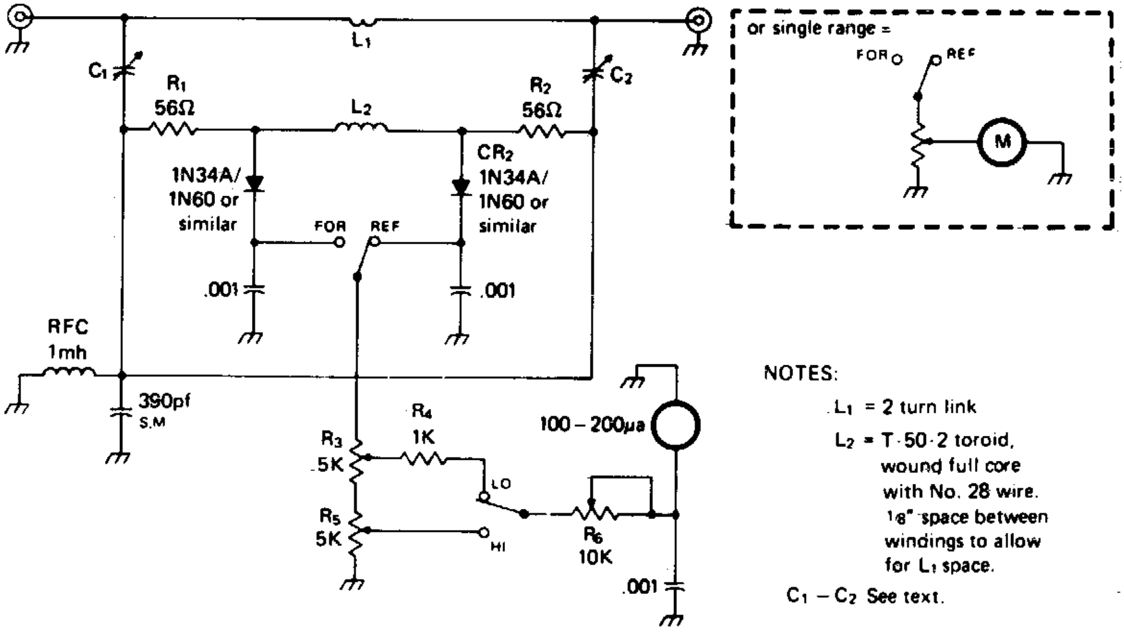

RG58 PASSED THROUGH TOROID
IN
SKT1
OUT
SKT2
T1
C1
10p
R2
150R
R3
150R
R5
1k
R6
470R
D3
OA95
RFC1
C2
40p
C3
100p
D1
OA95
D2
OA95
C6
1μ
C4
1μ
C5
1μ
R1
10k
R4
10k
R7
100k
Y
X
F
R
SW1A
SET
RV1
10k
F
R
POWER
SW1B
D4 & D5
1N914
M1
50μA F.S
2k

Fig. 2-2

Circuit Notes

Reflectometer (SWR Power Meter) covers three decades—from 100 kHz to 100 MHz. It can be constructed for rf powers as low as 500 mW or up to 500 watts.

IN-LINE WATTMETER


L1
C1
R1
56Ω
L2
R2
56Ω
C2
1N34A/
1N60 or
similar
CR2
1N34A/
1N60 or
similar
FOR
REF
.001
.001
RFC
1mh
390pf
S.M
R3
.5K
R4
1K
LO
HI
R5
5K
R6
10K
100 – 200μa
.001
or single range =
FOR
REF
M

NOTES:

L_1 = 2 turn link

L_2 = T-50-2 toroid, wound full core with No. 28 wire. 1/8" space between windings to allow for L_1 space.

$C_1 - C_2$ See text.

Fig. 2-3

Circuit Notes

The circuit is not frequency sensitive. Its calibration will be accurate over a wide frequency spectrum, such as the entire amateur hf spectrum, if the values of L2, the voltage divider capacitors C1-2 and C3, and the resistances of R1-2 are chosen properly. R1-2 and CR1-2 should be matched for best results. Generally, R1-2 must be small compared to the reactance of L2 so as to avoid any significant effect on the L2 current which is induced by the transmission line current flowing through L1. The lower frequency limit of the bridge is set by the R1-R2/Ls ratio, and the cutoff is at the point where the value of R1-R2 becomes significant with reference to the reactance of L2 at that frequency point.

CW SIGNAL PROCESSOR

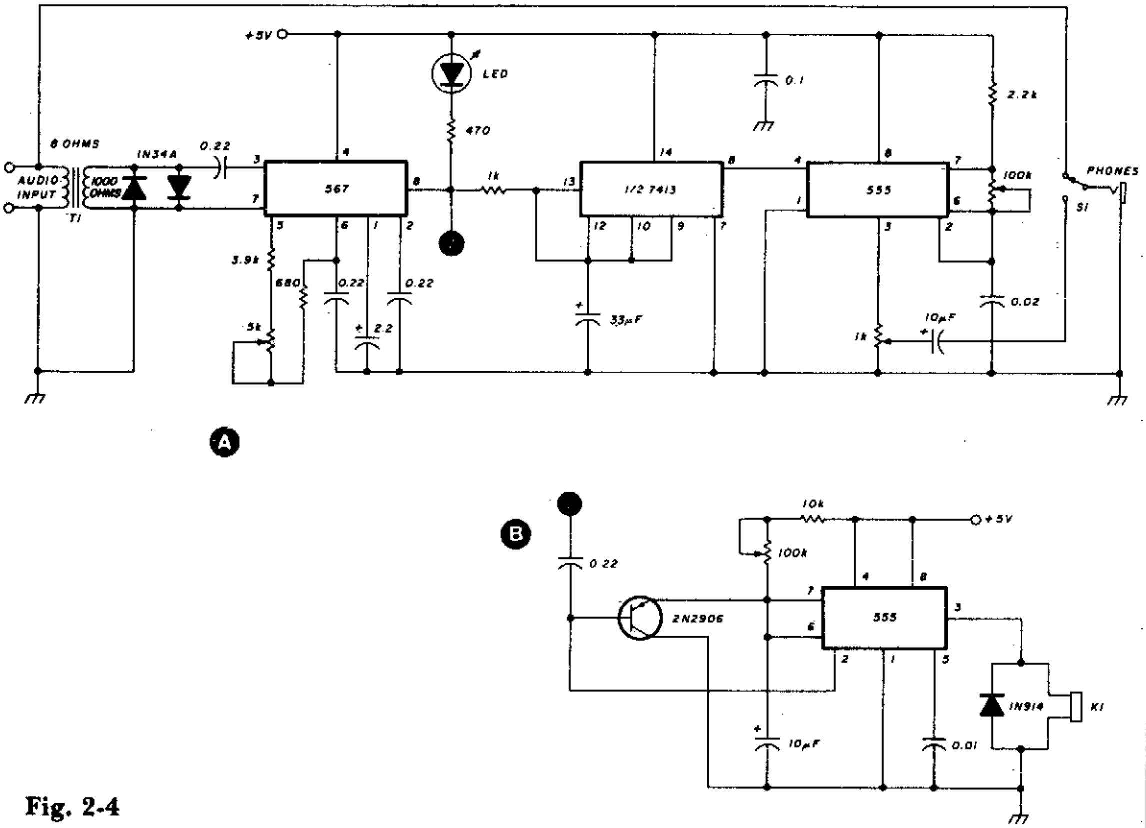
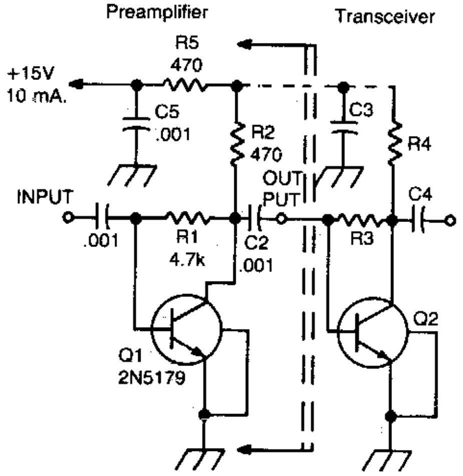

+5V
LED
470
0.1
2.2k
8 OHMS
1N34A
0.22
AUDIO INPUT
1000 OHMS
T1
567
1k
1/2 7413
555
100k
PHONES
S1
3.9k
680
0.22
0.22
5k
2.2
33μF
10μF
1k
0.02
A
B
10k
+5V
100k
0.22
2N2906
555
10μF
0.01
1N914
K1

Fig. 2-4

Circuit Notes

This circuit provides interferenced rejection for the CW operator. The 567 phase-locked loop is configured to respond to tones from 500 to 1100 Hz. The Schmitt trigger reduces the weighting effect caused by the output of the PLL remaining low after removal of the audio signal. Ten to 15 millivolts of audio activate the circuit. For periods of loss of signal, circuit B will automatically switch back to live receiver audio after a suitable delay. (If a relay with a 5-volt coil is not available, the circuit can also be powered from +12 volts.) When circuit B is used, the contacts on relay K1 replace S1.

TWO-METER PREAMPLIFIER FOR HANDITALKIES

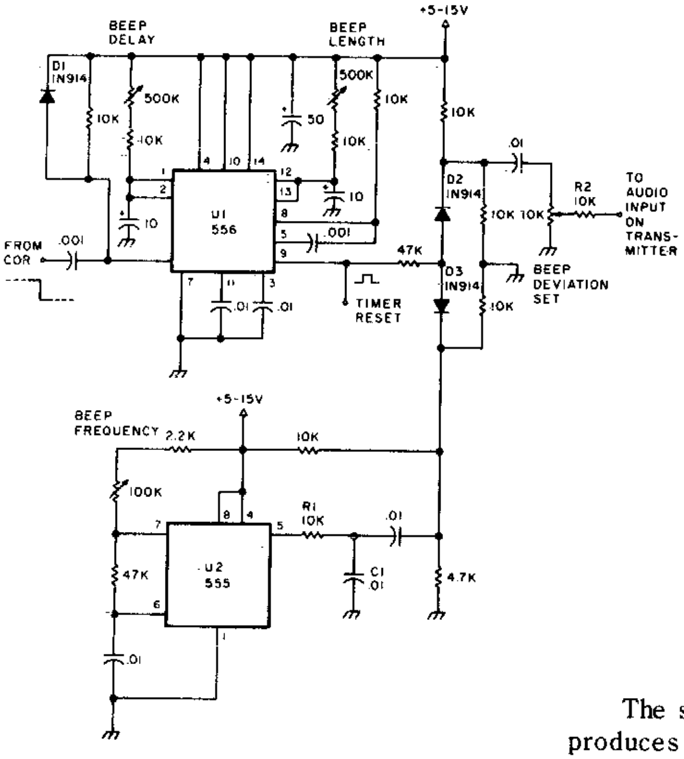

Preamplifier
Transceiver
+15V
10 mA.
R5
470
C5
.001
R2
470
C3
R4
INPUT
.001
R1
4.7k
C2
.001
OUT PUT
C4
R3
Q1
2N5179
Q2

Circuit Notes

This simple, inexpensive, wideband rf amplifier provides 14 dB gain on two meters without the use of tuned circuits.

Fig. 2-5

REPEATER BEEPER


+5-15V
BEEP DELAY
BEEP LENGTH
D1 IN914
500K
10K
10K
50
500K
10K
10K
10K
FROM COR
.001
10
U1 556
10
.001
47K
.01
D2 IN914
D3 IN914
10K
10K
R2 10K
TO AUDIO INPUT ON TRANS-MITTER
BEEP DEVIATION SET
10K
.01
.01
TIMER RESET
+5-15V
BEEP FREQUENCY
2.2K
10K
100K
R1 10K
.01
47K
U2 555
C1 .01
4.7K
.01

DELAY RANGE 0.15 TO 5 SECONDS
BURST RANGE 0.15 TO 5 SECONDS
TONE RANGE 500 TO 1400Hz

Fig. 2-6

Circuit Notes

The signal from COR triggers U1 which produces a beep-gate pulse that enables the analog gate consisting of D2 and D3 to pass the beep tone generated by U2.

ELECTRONIC KEYER

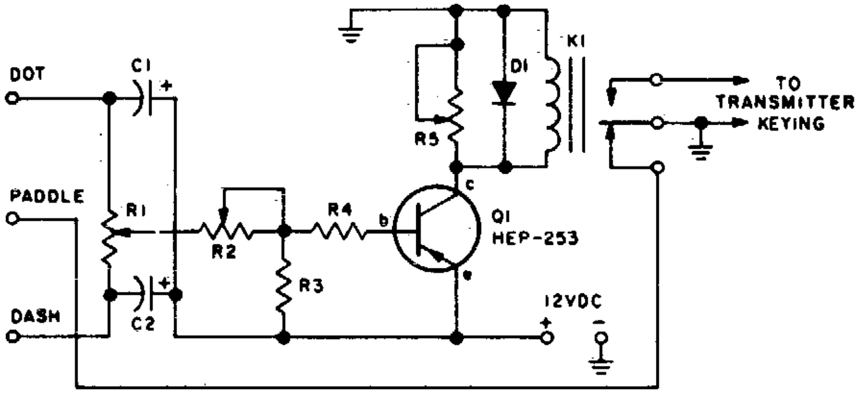

DOT
C1
PADDLE
R1
R2
R4
b
R5
D1
K1
c
Q1
HEP-253
e
R3
DASH
C2
12VDC
+
-
TO
TRANSMITTER
KEYING

Fig. 2-7

PARTS LIST FOR HAM'S KEYER
C1—3-uF, 6-VDC electrolytic capacitor
C2—10-uF, 6-VDC electrolytic capacitor
D1—1N60 diode
K1—12-VDC relay
Q1—HEP-253 pnp transistor
R1—10,000-ohm linear potentiometer
R2—50,000-ohm potentiometer
R3—1200-ohm, ½-watt resistor
R4—560-ohm, ½-watt resistor
R5—5000-ohm potentiometer

Circuit Notes

This circuit automatically produces Morse code dots and dashes set by time constants involving C1 and C2. R1 sets dot/dash ratio and R2 sets the speed. R5 sets the relay drop-out point.

CODE PRACTICE OSCILLATOR

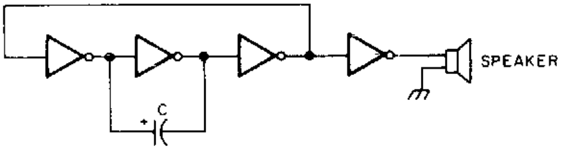
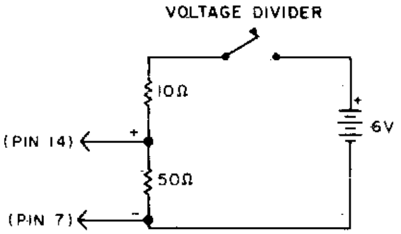
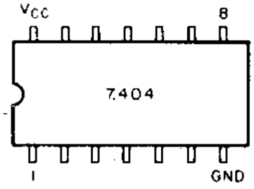


Fig. 2-8


VOLTAGE DIVIDER
10Ω
+
6V
(PIN 14)
+
50Ω
(PIN 7)
-


Vcc
8
7404
1
GND

Circuit Notes

This simple cpo uses the 7404 low-power Schottky hex inverter. C is a 5- to 30-μF electrolytic selected for the desired pitch. The speaker is a 2-inch, 8-ohm unit.

AUTOMATIC TAPE RECORDING

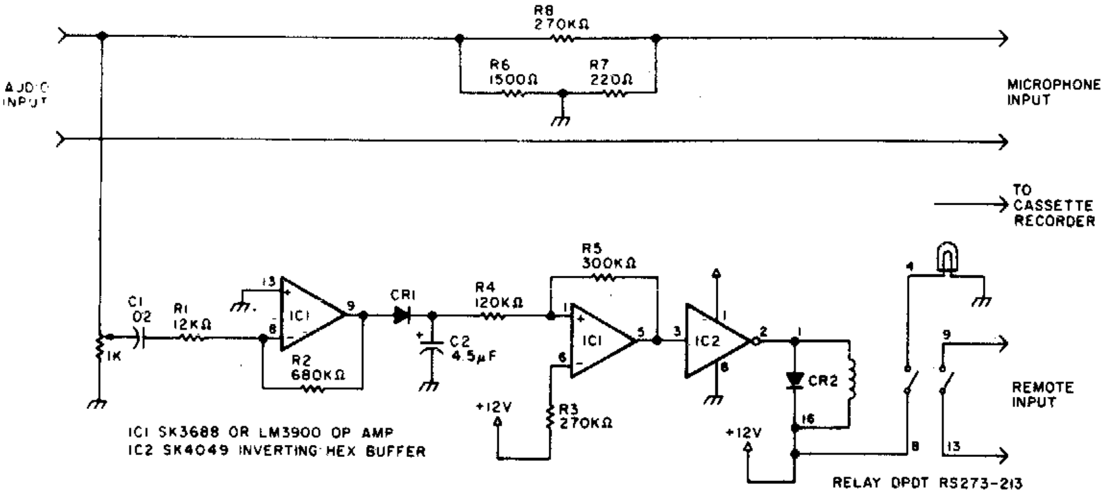

R8
270KΩ
R6
1500Ω
R7
220Ω
AUDIO INPUT
MICROPHONE INPUT
TO CASSETTE RECORDER
R5
300KΩ
C1
.02
R1
12KΩ
IC1
CR1
R4
120KΩ
IC1
IC2
CR2
1K
R2
680KΩ
C2
4.5μF
+12V
R3
270KΩ
+12V
REMOTE INPUT
IC1 SK3688 OR LM3900 OP AMP
IC2 SK4049 INVERTING HEX BUFFER
RELAY DPDT RS273-213

Fig. 2-9

Circuit Notes

Amateurs don't have to miss the action while away from the rig. This circuit turns on a tape recorder whenever the receiver's squelch is broken. After signal loss, the recorder will shut off following a slight delay.

SELF-POWERED CW MONITOR

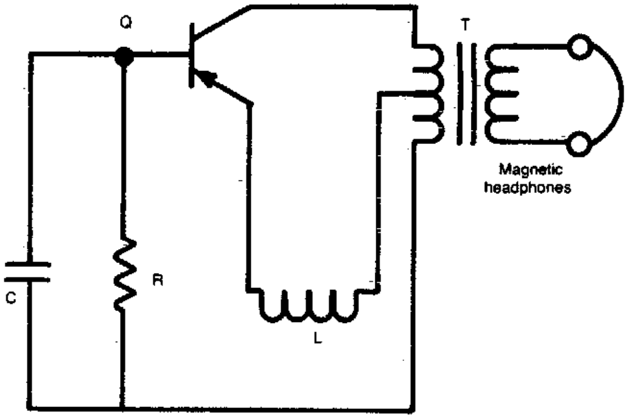
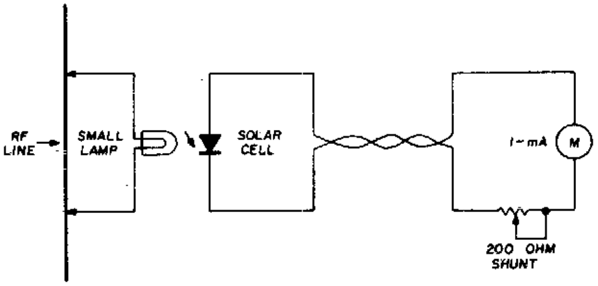

Q
T
Magnetic headphones
C
R
L

Fig. 2-10

Circuit Notes

Position L near the transmitter output tank to hear the key-down tone. Then tape the coil in place. C = .047 μF, R = 8.2 K, Q = HEP 253 (or equal), T = 500: 500 ohm center tapped transformer. L = 2 to 6 turns on ½″ coil form.

REMOTE RF CURRENT READOUT

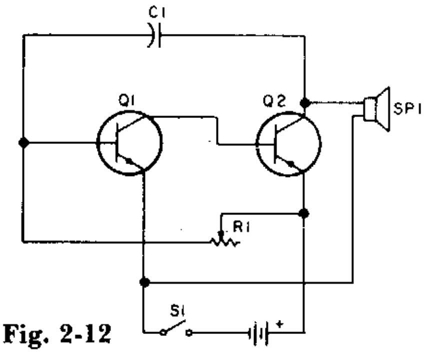

RF LINE
SMALL LAMP
SOLAR CELL
1-mA
M
200 OHM SHUNT

Circuit Notes

A suitable pilot lamp is illuminated by a small sample of rf and energizes an inexpensive solar cell; the dc current generated by the cell is a measure of relative rf power, and may be routed to a low-current meter located at any convenient point. A sensitive, low-current pilot lamp is desirable to cause minimum disturbance to normal rf circuit conditions. The number 48 or 49, 60 mA lamp is suitable for use with transmitters above 1-watt output.

Fig. 2-11

CODE PRACTICE OSCILLATOR

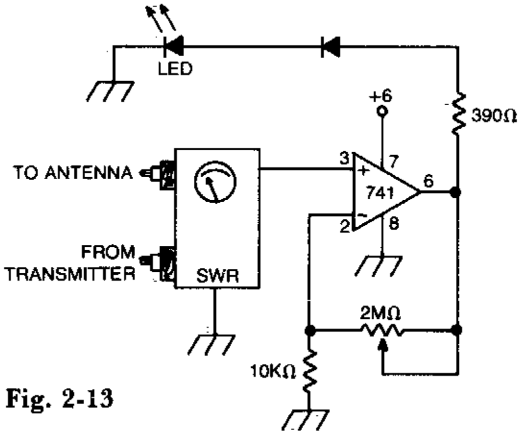

C1
Q1
Q2
SP1
R1
S1
+

Circuit Notes

Oscillator, works with 2 to 12 Vdc (but 9 to 12 volts gives best volume and clean keying). R1 can be replaced with a 500 K pot and the circuit will sweep the entire audio frequency range.

Fig. 2-12

SWR WARNING INDICATOR


LED
+6
390Ω
TO ANTENNA
FROM TRANSMITTER
SWR
3
7
+
741
6
2
−
8
2MΩ
10KΩ

Circuit Notes

Op amp with dc input from SWR meter can be adjusted to preset the SWR reading at which the LED lights.

Fig. 2-13

SUBAUDIBLE TONE ENCODER

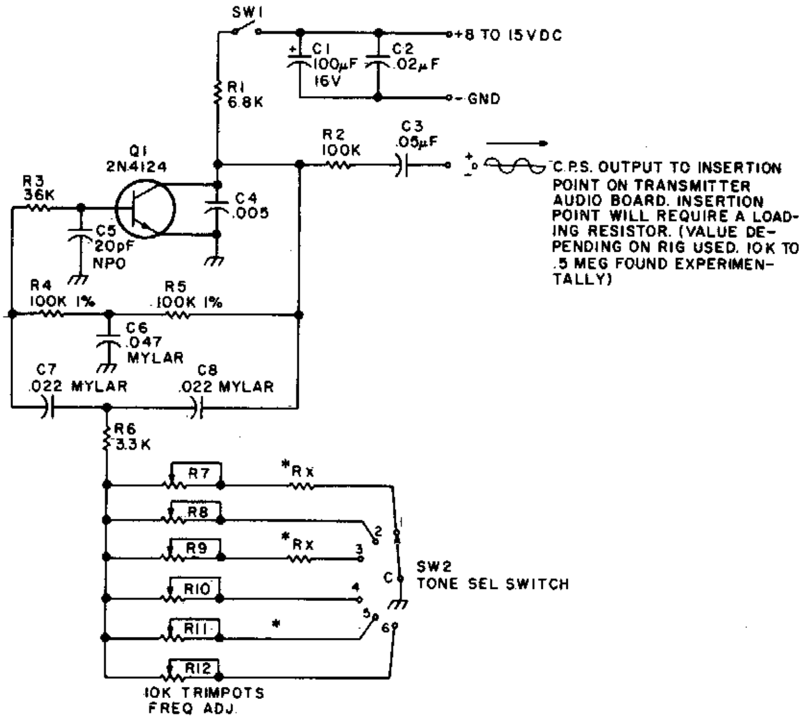

SW1
+8 TO 15VDC
C1 100µF 16V
C2 .02µF
R1 6.8K
GND
R2 100K
C3 .05µF
Q1 2N4124
R3 36K
C4 .005
C5 20pF NPO
C.P.S. OUTPUT TO INSERTION POINT ON TRANSMITTER AUDIO BOARD. INSERTION POINT WILL REQUIRE A LOADING RESISTOR. (VALUE DEPENDING ON RIG USED. 10K TO .5 MEG FOUND EXPERIMENTALLY)
R4 100K 1%
R5 100K 1%
C6 .047 MYLAR
C7 .022 MYLAR
C8 .022 MYLAR
R6 3.3K
R7
*Rx
R8
R9
*Rx
R10
R11
*
R12
1
2
3
4
5
6
C
SW2 TONE SEL SWITCH
10K TRIMPOTS FREQ ADJ.

Fig. 2-14

Circuit Notes

This twin-T oscillator produces six preset subaudible tones from 93 to 170 Hz in three ranges.

AUDIO MIXER

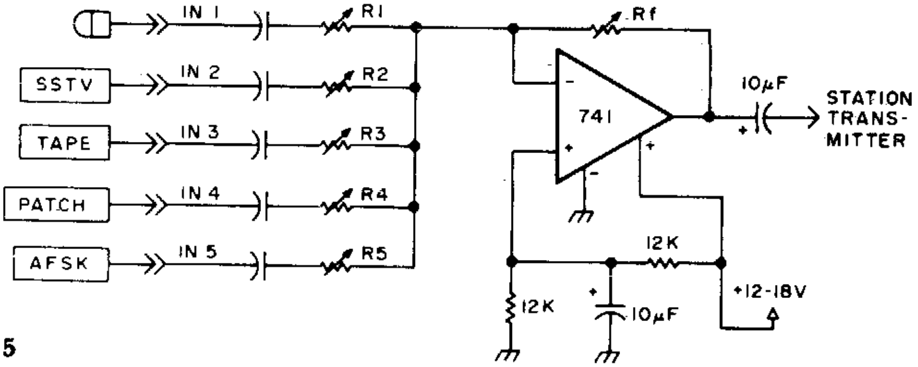
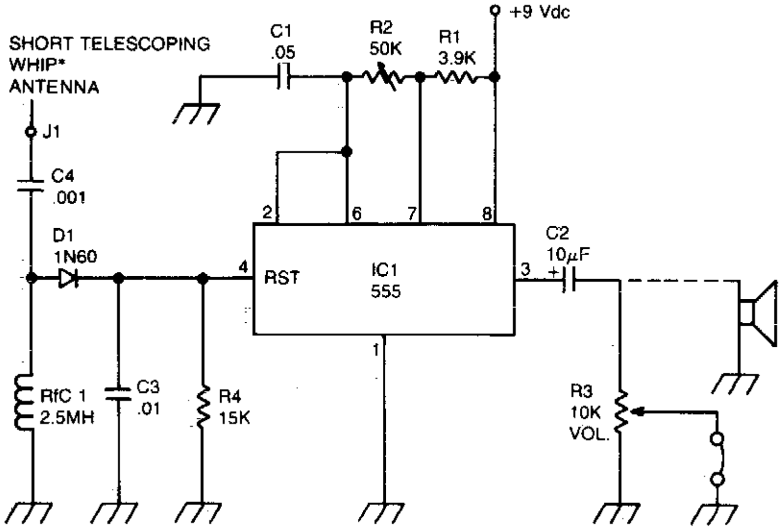

IN 1
SSTV
IN 2
TAPE
IN 3
PATCH
IN 4
AFSK
IN 5
R1
R2
R3
R4
R5
Rf
741
10µF
STATION TRANS-MITTER
12K
12K
10µF
+12-18V

Fig. 2-15

Circuit Notes

The 741 op amp is used as a summing amplifier to combine several audio inputs. Overall gain is set by R_f.

RF-POWERED SIDETONE OSCILLATOR


+9 Vdc
SHORT TELESCOPING WHIP* ANTENNA
C1 .05
R2 50K
R1 3.9K
J1
C4 .001
D1 1N60
2
6
7
8
4
RST
IC1 555
3
C2 10μF
+
1
RfC 1 2.5MH
C3 .01
R4 15K
R3 10K VOL.

*PORTABLE RADIO REPLACEMENT TYPE

Fig. 2-16

Circuit Notes

A sidetone oscillator is a special audio oscillator that is turned on and off with the transmitter. The oscillator is rf-driven and battery operated. It uses a 555 IC timer as an astable multivibrator. Keying is accomplished by applying a positive dc potential, developed from the rf signal, to the reset terminal of the 555.

HARMONIC GENERATOR

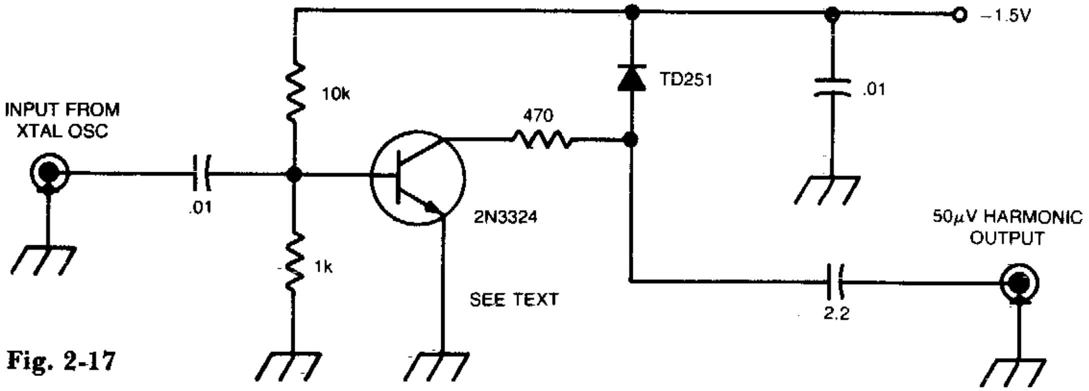

−1.5V
INPUT FROM XTAL OSC
10k
470
TD251
.01
.01
2N3324
1k
SEE TEXT
50μV HARMONIC OUTPUT
2.2

Fig. 2-17

Circuit Notes

This circuit will produce 50 μV harmonics through 1296 MHz with an input of 0.15-1 V from a 100 or 1000 kHz crystal oscillator. With a germanium diode instead of a tunnel diode, harmonics can be heard up to about 147 MHz.

AUTOMATIC TTL MORSE-CODE KEYER

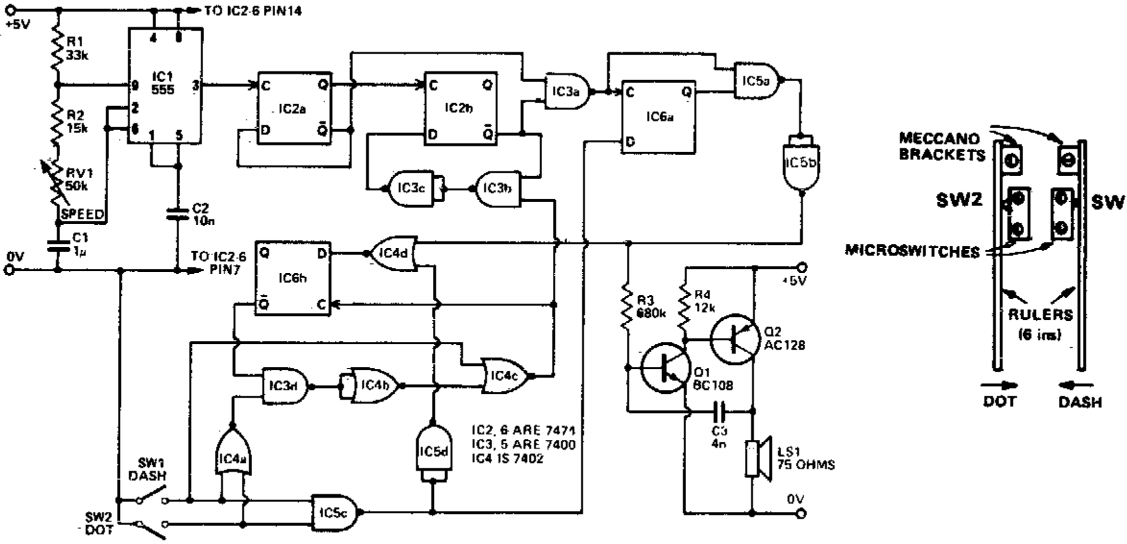

+5V
TO IC2-6 PIN14
R1
33k
IC1
555
R2
15k
RV1
50k
SPEED
C1
1μ
0V
C2
10n
TO IC2-6
PIN7
IC2a
IC2b
IC3a
IC5a
IC6a
IC5b
IC3c
IC3b
IC4d
IC6b
IC3d
IC4b
IC4c
IC4a
IC5d
IC5c
SW1
DASH
SW2
DOT
IC2, 6 ARE 7474
IC3, 5 ARE 7400
IC4 IS 7402
R3
680k
R4
12k
+5V
Q1
BC108
Q2
AC128
C3
4n
LS1
75 OHMS
0V
MECCANO
BRACKETS
SW2
SW1
MICROSWITCHES
RULERS
(6 ins)
DOT
DASH

Fig. 2-18

Circuit Notes

Automatically generated dits and dahs are produced over a speed range of 11 to 39 wpm. The upper limit can be raised by decreasing R2. SW1 and SW2 can be a "home-brew" paddle operated key.

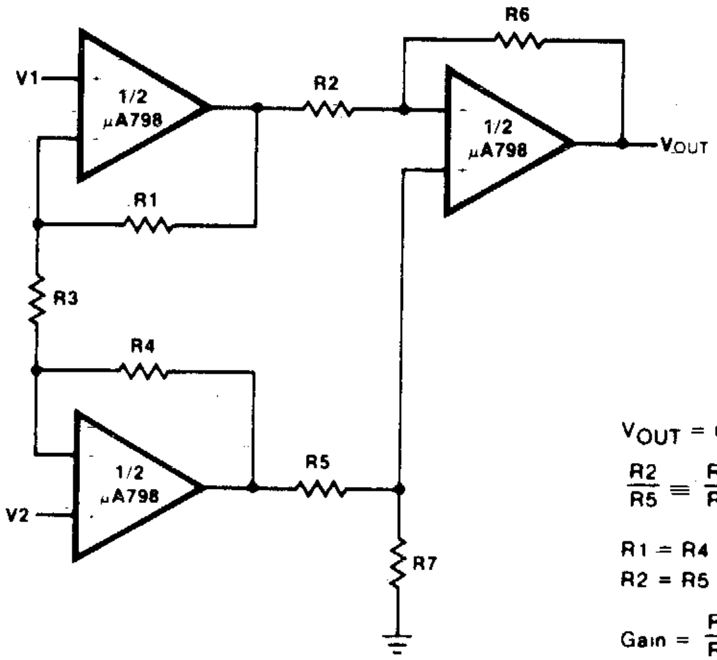
3

Amplifiers

The sources of the following circuits are contained in the Sources section beginning on page 730. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

High Impedance Differential Amplifier
Unity Gain Follower
Voltage Controlled Variable Gain Amplifier
Power Booster
Logarithmic Amplifier
Voltage Controlled Variable Gain Amplifier
Discrete Current Booster
Precision Process Control Interface
Voltage Controlled Amplifier
Absolute Value Amplifier
Programmable Gain Noninverting Amplifier with Selectable Inputs
× 1000 Amplifier Circuit
Inverting Amplifier with Balancing Circuit
Switching Power Amplifier
Precision Power Booster
Noninverting Voltage Follower
Color Video Amplifier
Fast Voltage Follower
Isolation Amplifier for Capacitive Loads
Cable Bootstrapping
Current Booster
Wideband Unity Gain Inverting Amplifier in a 75 Ohm System
High-Speed Current to Voltage Output Amplifier
Gated Amplifier
Reference Voltage Amplifier
Fast Summing Amplifier
Adjustment-Free Precision Summing Amplifier
Summing Amplifier with Low Input Current
× 10 Operational Amplifier Using L161
× 100 Operational Amplifier Using L161
Precision Absolute Value Circuit
Ultra-Low-Leakage Preamp
Dc to Video Log Amplifier
±100 V Common Mode Range Differential Amplifier
Wide Bandwidth, Low Noise, Low Drift Amplifier
Signal Distribution Amplifier
Audio Distribution Amplifier
High Input Impedance, High Output Current Voltage Follower
Precision Amplifier
Preamplifier and High-to-Low Impedance Converter
Noninverting Amplifier
High Impedance, High Gain, High Frequency Inverting Amp
Log-Ratio Amplifier
Inverting Amplifier
Logarithmic Amplifier

HIGH IMPEDANCE DIFFERENTIAL AMPLIFIER


R6
V1
1/2
μA798
R2
1/2
μA798
VOUT
R1
R3
R4
R5
1/2
μA798
V2
R7

$$V_{OUT} = C\,(1 + a + b)(V2 - V1)$$

$$\frac{R2}{R5} \equiv \frac{R6}{R7} \text{ for best CMRR}$$

$$R1 = R4$$

$$R2 = R5$$

$$\text{Gain} = \frac{R6}{R2}\left(1 + \frac{2R1}{R3}\right) = C\,(1 + a + b)$$

Fig. 3-1

UNITY GAIN FOLLOWER

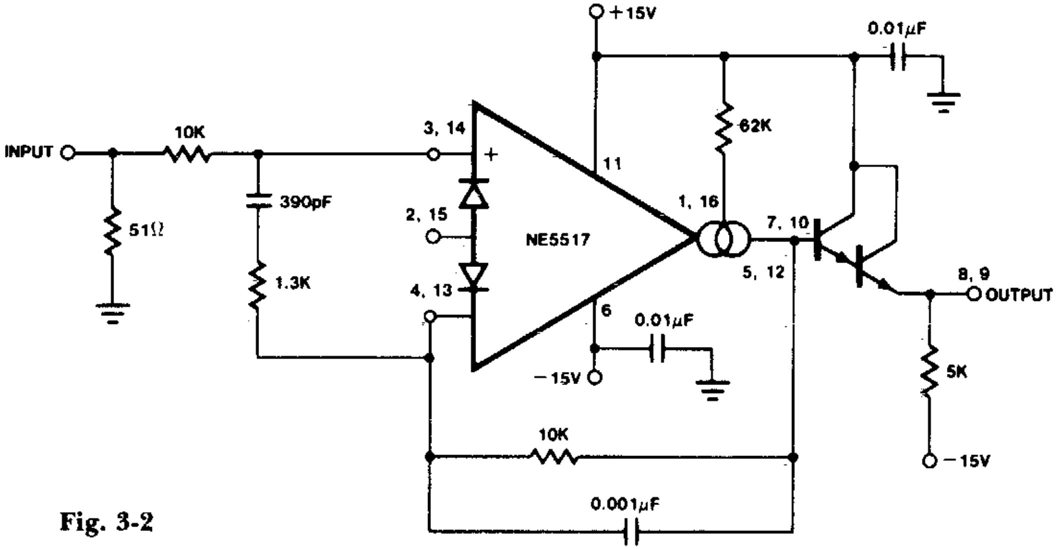

+15V
0.01μF
INPUT
10K
3, 14
+
11
62K
390pF
51Ω
2, 15
1, 16
7, 10
NE5517
5, 12
1.3K
4, 13
8, 9
OUTPUT
6
0.01μF
−15V
5K
10K
−15V
0.001μF

Fig. 3-2

VOLTAGE CONTROLLED VARIABLE GAIN AMPLIFIER

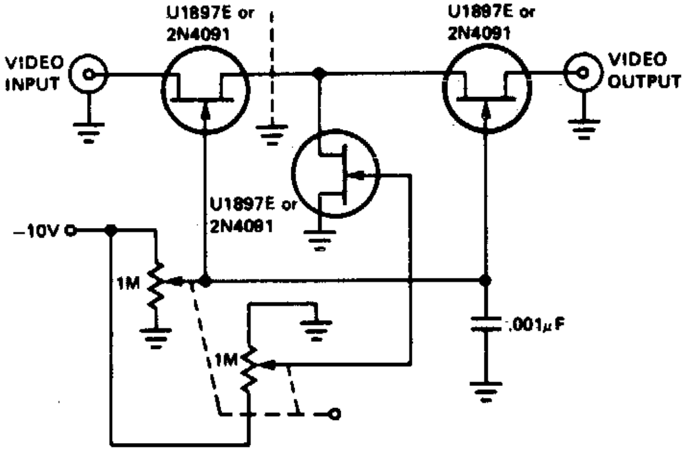
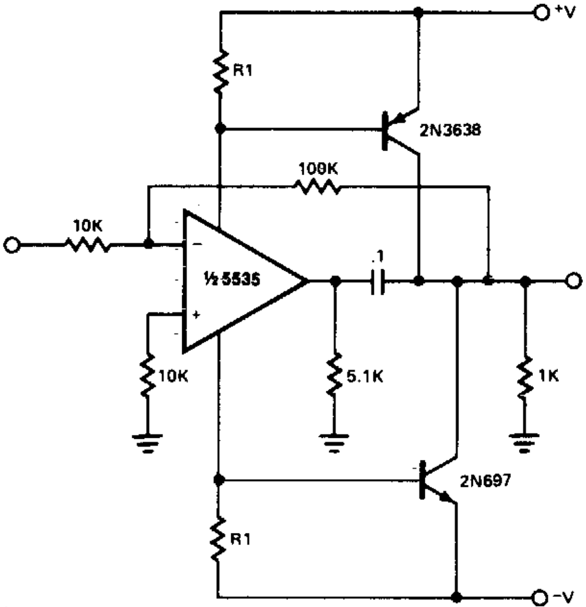

U1897E or
2N4091
U1897E or
2N4091
VIDEO
INPUT
VIDEO
OUTPUT
U1897E or
2N4091
−10V
1M
1M
.001μF

Fig. 3-3

Circuit Notes

The tee attenuator provides for optimum dynamic linear range attenuation up to 100 dB, even at f = 10.7 MHz with proper layout.

POWER BOOSTER


+V
R1
2N3638
100K
10K
−
+
½ 5535
.1
10K
5.1K
1K
2N697
R1
−V

All resistor values are in ohms.

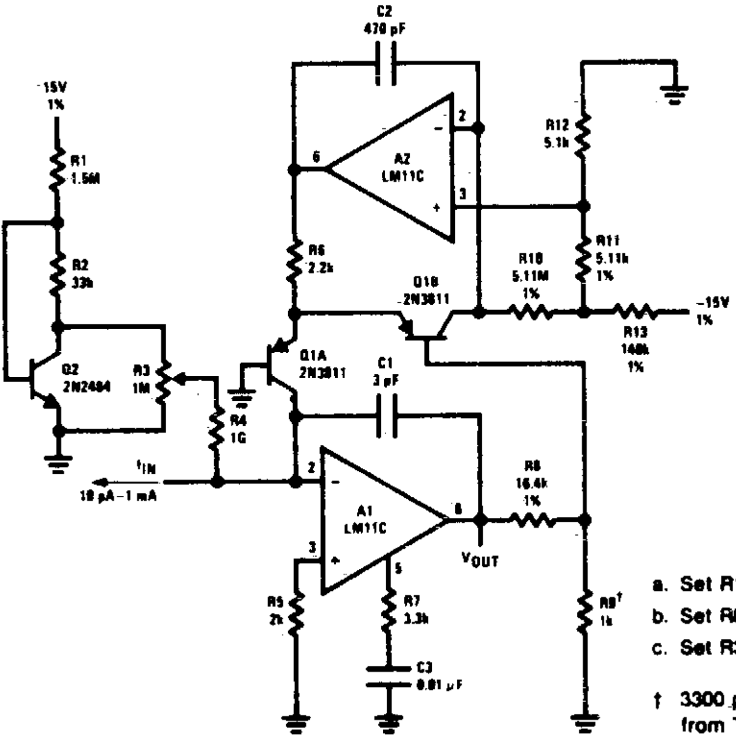
Fig. 3-4

Circuit Notes

Power booster is capable of driving moderate loads. The circuit as shown uses a NE5535 device. Other amplifiers may be substituted only if R1 values are changed because of the Icc current required by the amplifier. R1 should be calculated from the following expression:

$$R1 = \frac{600\text{ mW}}{I_{CC}}$$

LOGARITHMIC AMPLIFIER


C2
470 pF
-15V
1%
R1
1.5M
R2
33k
Q2
2N2484
R3
1M
R4
1G
A2
LM11C
6
2
3
R12
5.1k
R6
2.2k
Q1B
2N3811
R10
5.11M
1%
R11
5.11k
1%
-15V
1%
R13
140k
1%
Q1A
2N3811
C1
3 pF
I_{IN}
10 pA - 1 mA
2
A1
LM11C
6
R8
16.4k
1%
3
5
V_{OUT}
R5
2k
R7
3.3k
C3
0.01 μF
R9†
1k

a. Set R11 for $V_{OUT} = 0$ at $I_{IN} = 100\ \mu A$

b. Set R8 for $V_{OUT} = 3V$ at $I_{IN} = 100\ \mu A$

c. Set R3 for $V_{OUT} = -4V$ at $I_{IN} = 10\ pA$

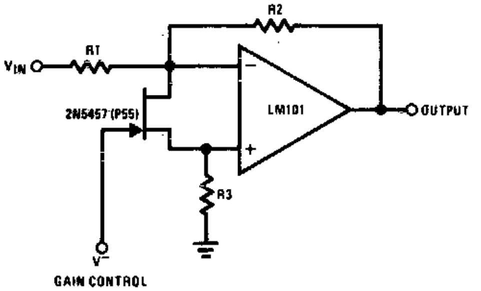
† 3300 ppM/°C. Type Q209 available from Tel Labs, Inc., Manchester, N.H.

Fig. 3-5

Circuit Notes

Unusual frequency compensation gives this logarithmic converter a 100 μs time constant from 1 mA down to 100 μA, increasing from 200 μs to 200 ms from 10 nA to 10 pA. Optional bias current compensation can give 10 pA resolution from – 55 °C to 100 °C. Scale factor is 1 V/decade and temperature compensated.

VOLTAGE CONTROLLED VARIABLE GAIN AMPLIFIER


R2
R1
V_{IN}
2N5457 (P55)
LM101
OUTPUT
R3
V−
GAIN CONTROL

Circuit Notes

The 2N5457 acts as a voltage variable resistor with an $R_{ds(on)}$ of 800 ohms max. Since the differential voltage on the LM101 is in the low mV range, the 2N5457 JFET will have linear resistance over several decades of resistance providing an excellent electronic gain control.

Fig. 3-6

DISCRETE CURRENT BOOSTER

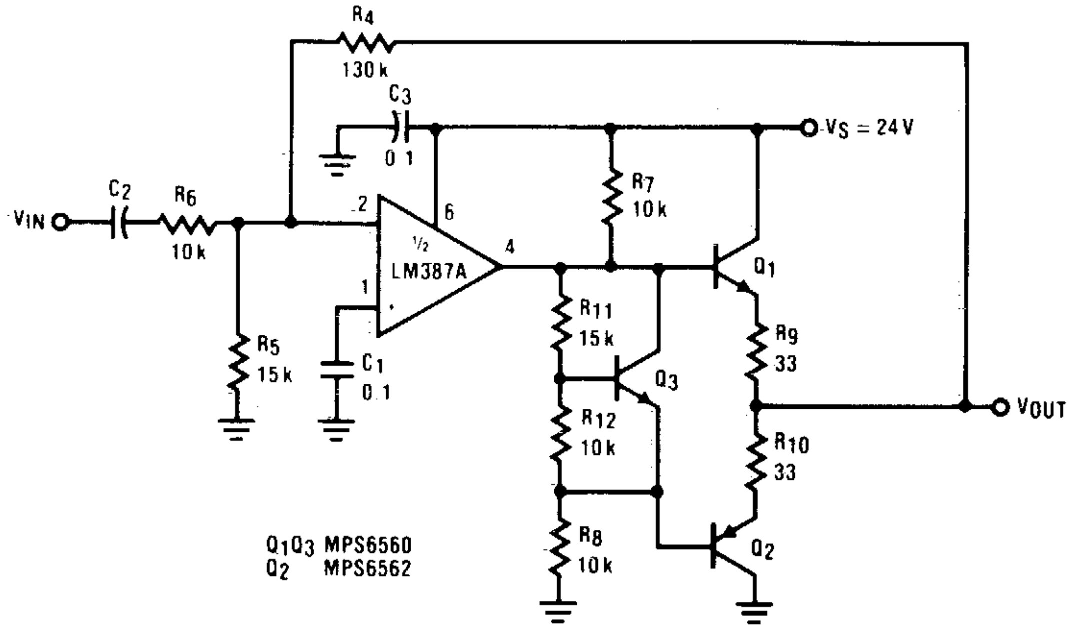

R4
130 k
C3
0 1
VS = 24 V
C2
R6
VIN
10 k
2
6
½
LM387A
4
1
R7
10 k
Q1
R11
15 k
R9
33
R5
15 k
C1
0 1
Q3
R12
10 k
VOUT
R10
33
R8
10 k
Q2
Q1Q3 MPS6560
Q2 MPS6562

Fig. 3-7

PRECISION PROCESS CONTROL INTERFACE

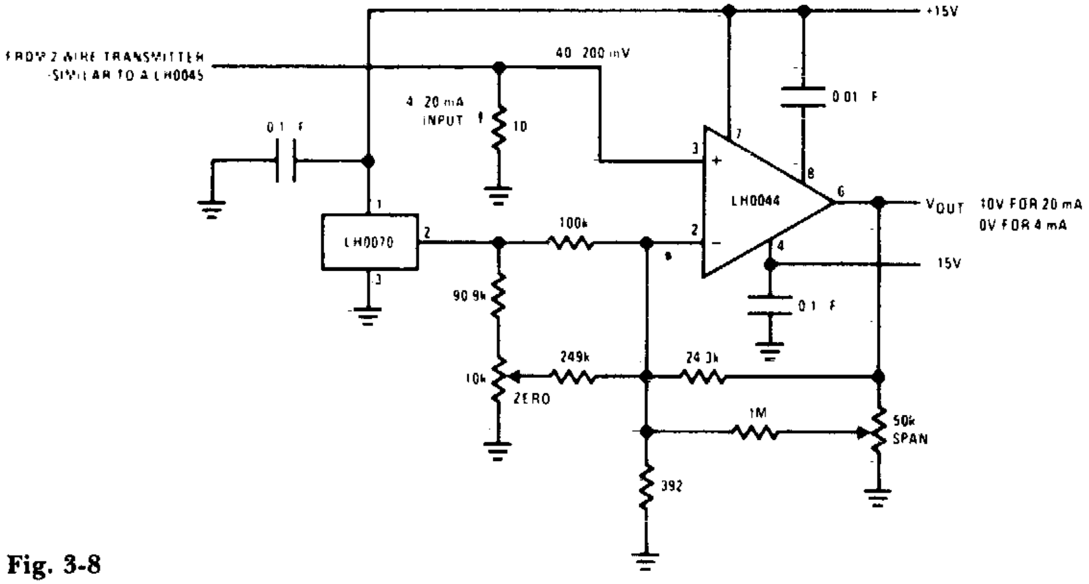
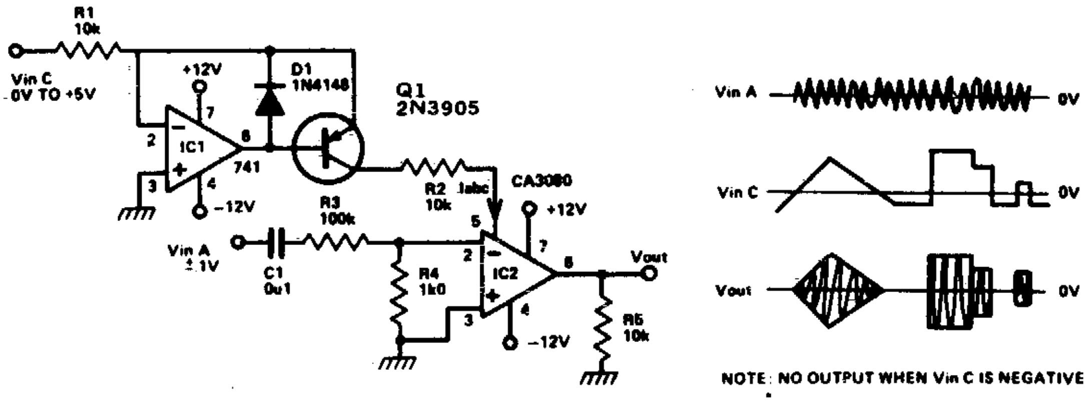

FROM 2 WIRE TRANSMITTER
SIMILAR TO A LH0045
40 200 mV
+15V
4 20 mA
INPUT
10
0 1 F
0 01 F
7
3
+
8
LH0044
6
VOUT
10V FOR 20 mA
0V FOR 4 mA
1
LH0070
2
3
100k
2
−
4
15V
0 1 F
90 9k
249k
24 3k
10k
ZERO
1M
50k
SPAN
392

Fig. 3-8

VOLTAGE CONTROLLED AMPLIFIER



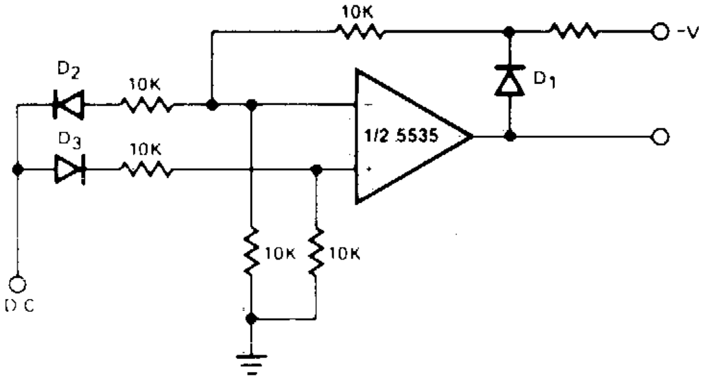
R1
10k
Vin C
0V TO +5V
+12V
D1
1N4148
Q1
2N3905
IC1
741
2
3
6
7
4
−12V
R3
100k
R2
10k
Iabc
CA3080
+12V
5
Vin A
±1V
C1
0u1
R4
1k0
IC2
2
3
7
6
4
−12V
Vout
R5
10k
Vin A
0V
Vin C
0V
Vout
0V
NOTE: NO OUTPUT WHEN Vin C IS NEGATIVE

Fig. 3-9

Circuit Notes

This circuit is basically an op amp with an extra input at pin 5. A current IABC is injected into this input and this controls the gain of the device linerly. Thus by inserting an audio signal (±10 mV) between pin 2 and 3 and by controlling the current on pin 5, the level of the signal output (pin 6) is controlled.

ABSOLUTE VALUE AMPLIFIER



10K
-V
D2
10K
D1
D3
10K
1/2 5535
10K
10K
DC

Fig. 3-10

Circuit Notes

The circuit generates a positive output voltage for either polarity of input. For positive signals, it acts as a noninverting amplifier and for negative signals, as an inverting amplifier. The accuracy is poor for input voltages under 1 V, but for less stringent applications, it can be effective.

PROGRAMMABLE GAIN NONINVERTING AMPLIFIER WITH SELECTABLE INPUTS

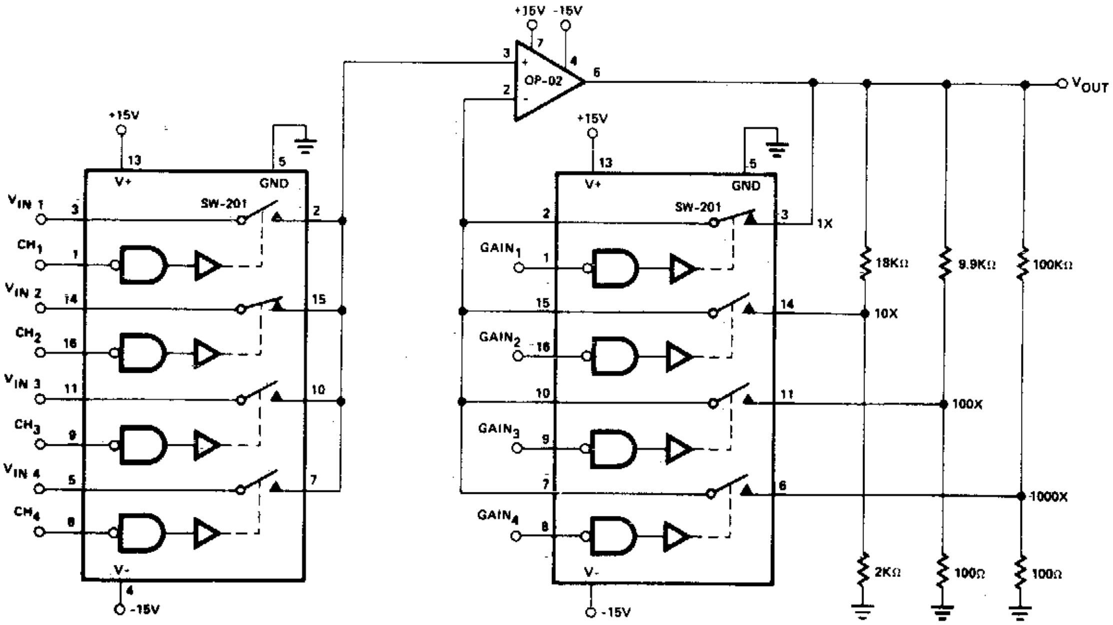
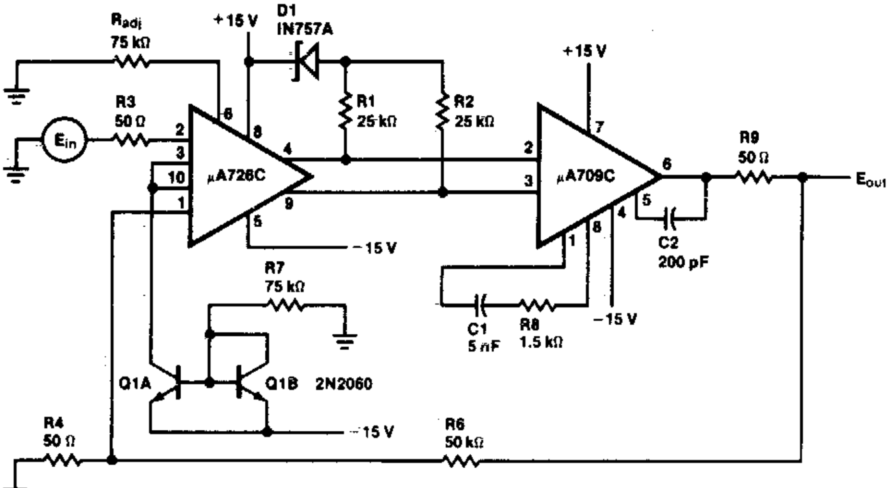

+15V
-15V
OP-02
VOUT
SW-201
V+
GND
V-
VIN 1
CH1
VIN 2
CH2
VIN 3
CH3
VIN 4
CH4
GAIN1
GAIN2
GAIN3
GAIN4
1X
10X
100X
1000X
18KΩ
9.9KΩ
100KΩ
2KΩ
100Ω
100Ω

Fig. 3-11

× 1000 AMPLIFIER CIRCUIT


Radj
75 kΩ
+15 V
D1
IN757A
R3
50 Ω
Ein
R1
25 kΩ
R2
25 kΩ
µA726C
µA709C
R9
50 Ω
Eout
C2
200 pF
-15 V
R7
75 kΩ
C1
5 nF
R8
1.5 kΩ
Q1A
Q1B
2N2060
R4
50 Ω
R6
50 kΩ

ALL RESISTORS 1%

Fig. 3-12

INVERTING AMPLIFIER WITH BALANCING CIRCUIT

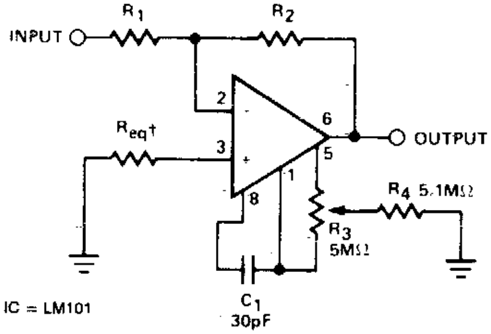

INPUT
R1
R2
Req†
2
3
6
5
1
8
OUTPUT
R4 5.1MΩ
R3
5MΩ
IC = LM101
C1
30pF

Fig. 3-13

Circuit Notes

Req may be zero or equal to the parallel combination of R1 and R2 for minimum offset.

PRECISION POWER BOOSTER

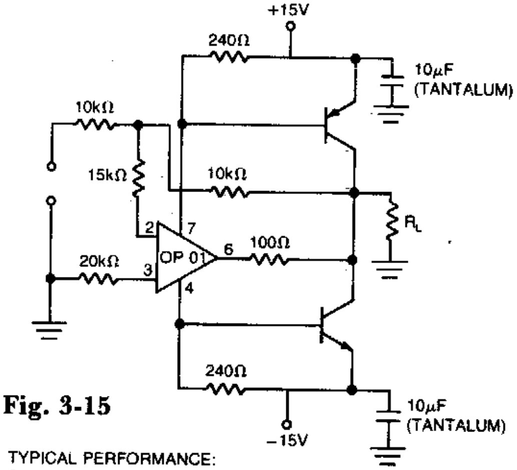

+15V
240Ω
10μF
(TANTALUM)
10kΩ
15kΩ
10kΩ
RL
2
7
OP 01
6
100Ω
20kΩ
3
4
240Ω
−15V
10μF
(TANTALUM)

Fig. 3-15

TYPICAL PERFORMANCE:
SLEW RATE------------≈18V/μSEC
0.1% SETTLING-------4μSEC (R_L = 500)
QUIESCENT SUPPLY CURRENT---1.5mA

SWITCHING POWER AMPLIFIER

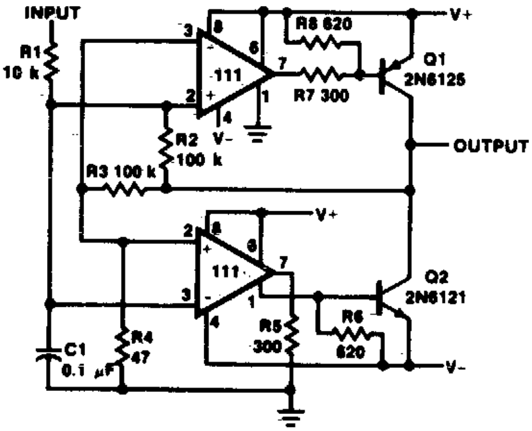
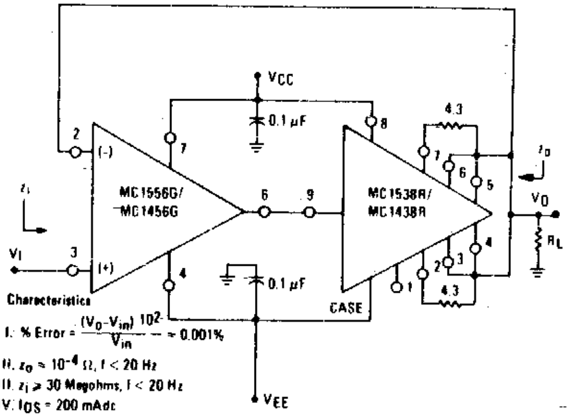

INPUT
R1
10 k
3
8
6
111
7
2
1
4
V−
R8 620
V+
Q1
2N6125
R7 300
OUTPUT
R2
100 k
R3 100 k
V+
2
8
6
111
7
3
1
4
Q2
2N6121
R5
300
R6
620
V−
C1
0.1 μF
R4
47

Fig. 3-14

NONINVERTING VOLTAGE FOLLOWER


VCC
0.1 μF
2
(−)
7
8
4.3
I_O
MC1556G/
MC1456G
6
9
MC1538R/
MC1438R
7
6
5
VO
RL
z_i
VI
3
(+)
4
0.1 μF
CASE
1
2
3
4
4.3
VEE

Characteristics

I. $\% \text{ Error} = \frac{(V_O - V_{in})\ 10^2}{V_{in}} \approx 0.001\%$

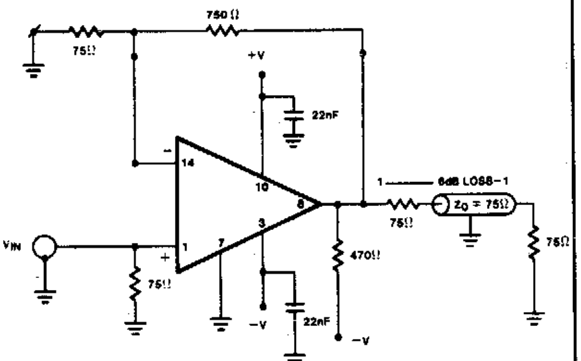
II. $z_O \approx 10^{-4}\ \Omega$, $f < 20$ Hz

III. $z_i \geq 30$ Megohms, $f < 20$ Hz

IV. I_{OS} = 200 mAdc

Fig. 3-16

COLOR VIDEO AMPLIFIER


750 Ω
75Ω
+V
22nF
14
10
8
1 ——— 6dB LOSS−1
Z_0 = 75Ω
75Ω
75Ω
VIN
1
7
3
470Ω
75Ω
−V
22nF
−V

IC = NE5539

Fig. 3-17

FAST VOLTAGE FOLLOWER

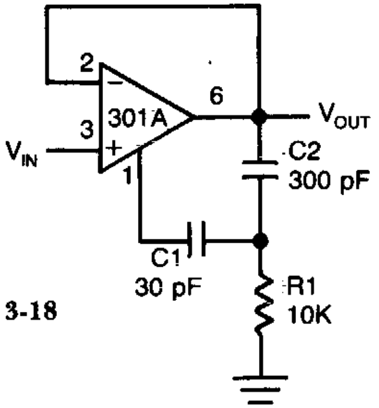
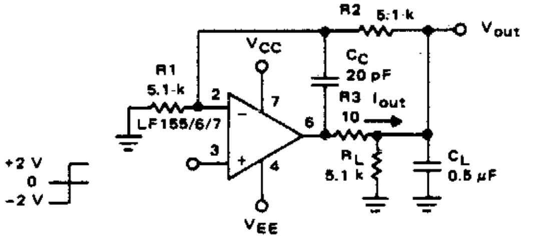

2
6
301A
3
VOUT
VIN
1
C2
300 pF
C1
30 pF
R1
10K

Fig. 3-18

Power Bandwidth: 15 kHz
Slew Rate: 1V/µs

ISOLATION AMPLIFIER FOR CAPACITIVE LOADS


R2 5.1 k
Vout
VCC
R1
5.1 k
CC
20 pF
2
7
R3 Iout
10
LF155/6/7
6
+2 V
0
−2 V
3
4
RL
5.1 k
CL
0.5 µF
VEE

- Overshoot 6%
- $t_s = 10\ \mu s$
- When driving large C_L, the V_{out} slew rate is determined by C_L and $I_{out(max)}$:

$$\frac{\Delta V_{out}}{\Delta t} = \frac{I_{out}}{C_L} \cong \frac{0.02}{0.5}\ V/\mu s = 0.04\ V/\mu s \text{ (with } C_L \text{ shown)}$$

Fig. 3-19

CABLE BOOTSTRAPPING

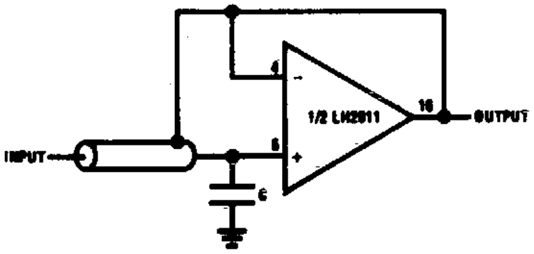
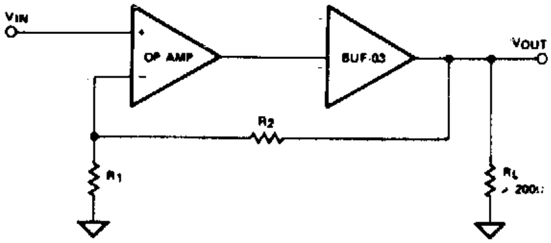

4
−
1/2 LH2011
16
OUTPUT
INPUT
5
+
6

Fig. 3-20

Circuit Notes

Bootstrapping input shield for a follower reduces cable capacitance, leakage, and spurious voltages from cable flexing. Instability can be avoided with small capacitor on input.

CURRENT BOOSTER



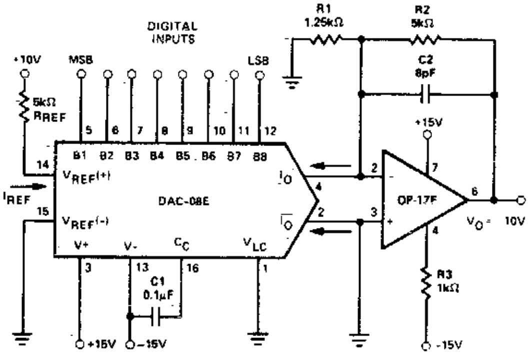
V_{IN}
OP AMP
BUF-03
V_{OUT}
R_2
R_1
R_L
≥ 200Ω

$\text{GAIN} = \frac{V_{OUT}}{V_{IN}} = 1 + \frac{R_2}{R_1}$

MAXIMUM LOAD CURRENT = ±50mA (10V ÷ 200Ω)

Fig. 3-21

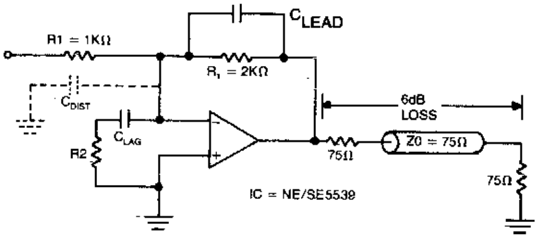
HIGH-SPEED CURRENT TO VOLTAGE OUTPUT AMPLIFIER



DIGITAL INPUTS
+10V
MSB
LSB
5kΩ
RREF
5
6
7
8
9
10
11
12
B1
B2
B3
B4
B5
B6
B7
B8
14
$V_{REF}(+)$
I_{REF}
15
$V_{REF}(-)$
DAC-08E
V+
V-
C_C
V_{LC}
I_O
$\overline{I_O}$
4
2
3
13
16
1
C1
0.1μF
+15V
-15V
R1
1.25kΩ
R2
5kΩ
C2
8pF
+15V
2
3
7
6
4
OP-17F
V_O = 10V
R3
1kΩ
-15V

Fig. 3-23

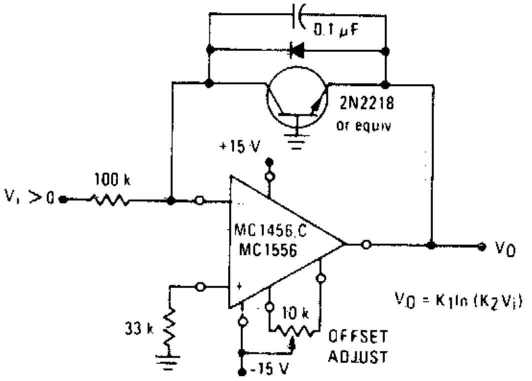
WIDEBAND UNITY GAIN INVERTING AMPLIFIER IN A 75 OHM SYSTEM



C_{LEAD}
R1 = 1KΩ
R_f = 2KΩ
C_{DIST}
C_{LAG}
R2
6dB LOSS
75Ω
Z0 = 75Ω
75Ω
IC = NE/SE5539

Fig. 3-22

LOGARITHMIC AMPLIFIER



0.1 μF
2N2218 or equiv
+15 V
100 k
$V_i > 0$
MC1456,C
MC1556
V_O
$V_O = K_1 \ln (K_2 V_i)$
33 k
10 k
OFFSET ADJUST
-15 V

Fig. 3-24

GATED AMPLIFIER

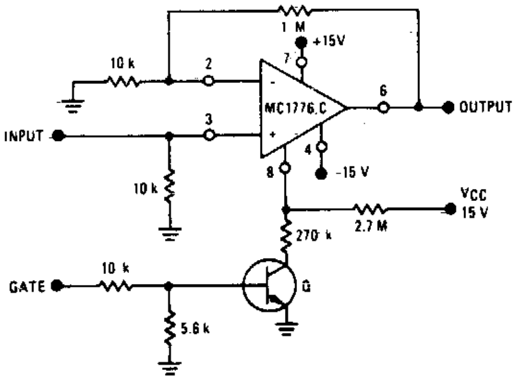
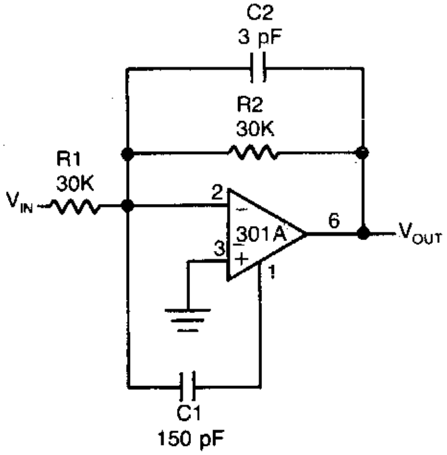

1 M
+15V
7
10 k
2
6
MC1776C
OUTPUT
INPUT
3
4
8
-15 V
10 k
VCC
15 V
270 k
2.7 M
10 k
GATE
Q
5.6 k

Fig. 3-25

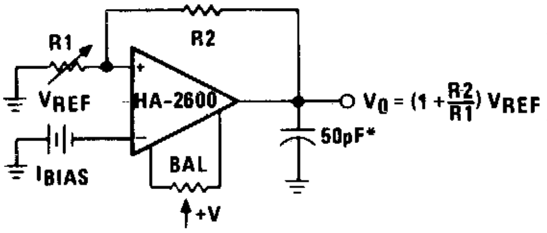
FAST SUMMING AMPLIFIER


C2
3 pF
R2
30K
R1
30K
V_{IN}
2
6
301A
V_{OUT}
3
1
C1
150 pF

Power Bandwidth: 250 kHz
Small Signal Bandwidth: 3.5 MHz
Slew Rate: 10V/μs

Fig. 3-27

REFERENCE VOLTAGE AMPLIFIER


R1
R2
VREF
HA-2600
$V_0 = (1 + \frac{R2}{R1})$ VREF
50pF*
IBIAS
BAL
+V

FEATURES

1 MINIMUM BIAS CURRENT IN REFERENCE CELL
2 SHORT CIRCUIT PROTECTION

IC = HA-OP07

Fig. 3-26

ADJUSTMENT-FREE PRECISION SUMMING AMPLIFIER

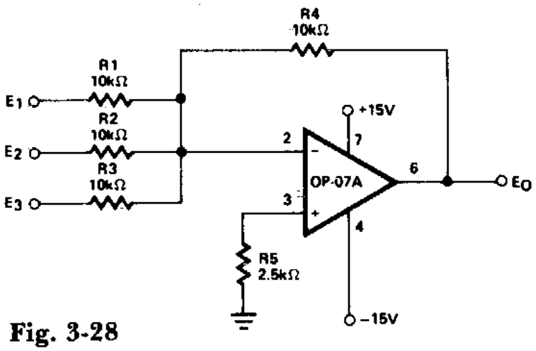
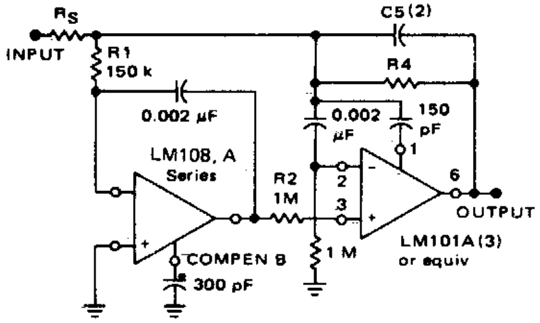

R4
10kΩ
R1
10kΩ
E1
R2
10kΩ
E2
R3
10kΩ
E3
+15V
7
2
6
OP-07A
EO
3
4
R5
2.5kΩ
−15V

Fig. 3-28

Circuit Notes

This circuit produces continuous outputs that are a function of multiple input variables.

SUMMING AMPLIFIER WITH LOW INPUT CURRENT


RS
INPUT
R1
150 k
C5 (2)
R4
0.002 μF
0.002 μF
150 pF
1
LM108, A Series
R2
1M
2
3
6
OUTPUT
COMPEN B
300 pF
1 M
LM101A (3) or equiv

(1) Power Bandwidth: 250 kHz
Small Signal Bandwidth: 3.5 MHz
Slew Rate: 10 V/μs

(2) $C5 = \frac{6 \times 10^{-8}}{R1}$

(3) In addition to increasing speed, the LM101A raises high and low frequency gain, increases output drive capability and eliminates thermal feedback.

Fig. 3-29

× 100 OPERATIONAL AMPLIFIER USING L161

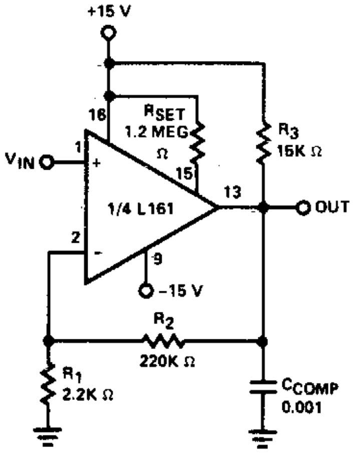

+15 V
16
RSET 1.2 MEG Ω
15
VIN
1
R3 15K Ω
1/4 L161
13
OUT
2
9
−15 V
R2
220K Ω
R1 2.2K Ω
CCOMP 0.001

Fig. 3-31

Circuit Notes

Amplifier has gain-bandwidth product of 20 MHz with slew rate of 0.3V/μ sec.

× 10 OPERATIONAL AMPLIFIER USING L161

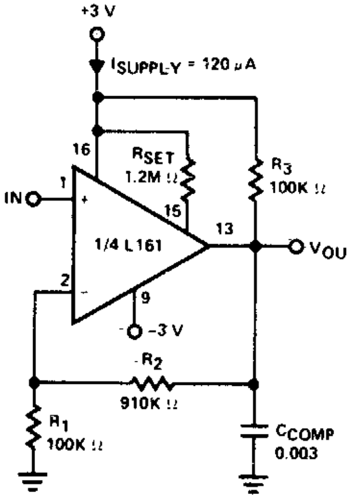
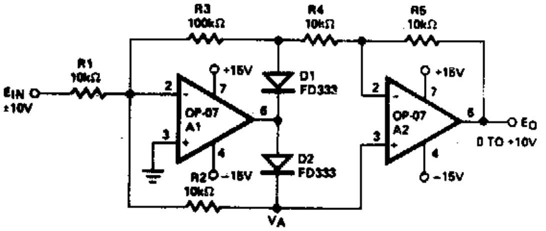

+3 V
ISUPPLY = 120 μA
16
RSET 1.2M Ω
15
IN
1
R3 100K Ω
1/4 L161
13
VOUT
2
9
−3 V
R2
910K Ω
R1 100K Ω
CCOMP 0.003

Fig. 3-30

Circuit Notes

Amplifier is 3 dB down at 100 kHz and has a slew rate of 0.02V/μ sec.

PRECISION ABSOLUTE VALUE CIRCUIT


R3 100kΩ
R4 10kΩ
R5 10kΩ
R1 10kΩ
EIN ±10V
+15V
D1 FD333
2
7
OP-07 A1
6
3
4
R2 10kΩ
−15V
D2 FD333
VA
+15V
2
7
OP-07 A2
6
3
4
−15V
EO
0 TO +10V

POSITIVE INPUT

1. $V_A = 0$, D2 OFF, D1 ON
2. $E_O = \left(\frac{-E_{IN}\,R3}{R1}\right) \cdot \left(\frac{-R5}{R4}\right) = E_{IN}\frac{R3\,R5}{R1\,R4}$
3. WITH $R1 = R3 = R4 = R5$: $E_O = E_{IN}$
4. V_{OS} ERROR INCLUDED:
   $E_O = E_{IN} + 2V_{OS2}$

NEGATIVE INPUT

1. D1 OFF, D2 ON
2. $\frac{-E_{IN}}{R1} = \frac{V_A}{R2} + \frac{V_A}{R3 + R4}$
3. $E_O = V_A\left(1 + \frac{R5}{R3 + R4}\right)$
4. WITH $R3 = R4 = R5$: $E_O = 1.5V_A$
5. $E_O = -\frac{(R2)\,(R3 + R4)\,(1.5)\,E_{IN}}{R1\,(R2 + R3 + R4)}$
6. WITH $R1 = R2 = R3 = R4$: $E_O = -E_{IN}$
7. V_{OS} ERROR INCLUDED:
   $E_O = -E_{IN} + 1.5V_{OS2} - 0.5V_{OS1}$
8. FOR BOTH INPUTS: $E_O = + E_{IN}$

Fig. 3-32

ULTRA-LOW-LEAKAGE PREAMP

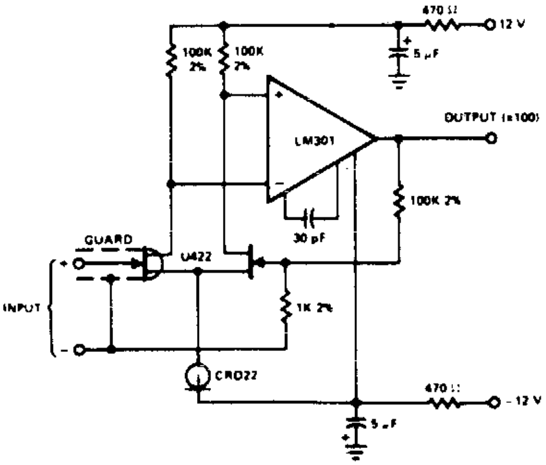

470 Ω
12 V
100K 2%
100K 2%
5 µF
LM301
OUTPUT (×100)
30 pF
100K 2%
GUARD
U422
INPUT
1K 2%
CR022
470 Ω
−12 V
5 µF

Fig. 3-33

Circuit Notes

Input leakage—2 pA at 75 °C.

DC TO VIDEO LOG AMPLIFIER

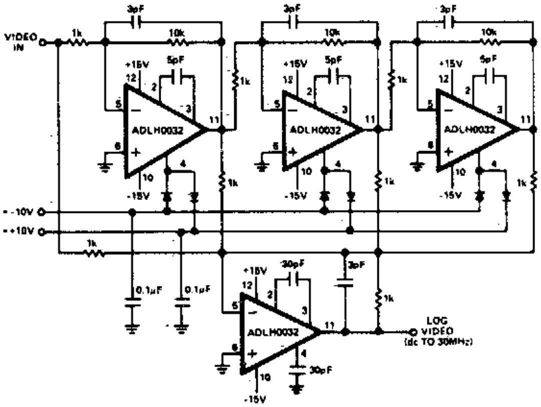
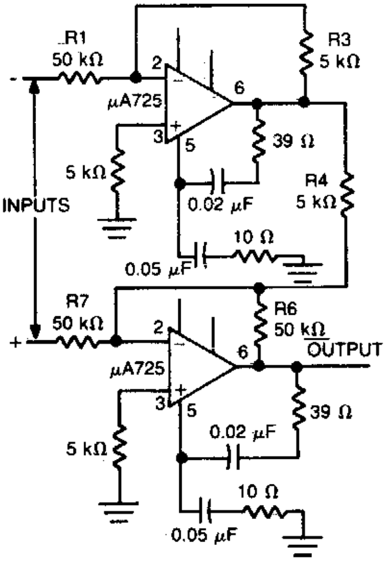

VIDEO IN
1k
3pF
10k
+15V
5pF
12
2
5
3
11
6
ADLH0032
10
4
−15V
1k
V_CL = −10V
V_CL = +10V
0.1µF
0.1µF
30pF
3pF
LOG VIDEO (dc TO 30MHz)
30pF

Fig. 3-34

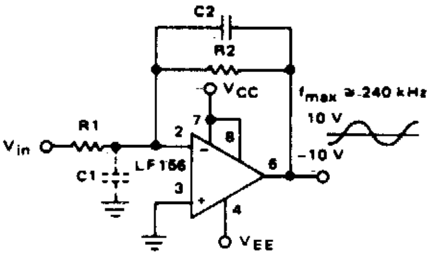
±100 V COMMON MODE RANGE DIFFERENTIAL AMPLIFIER


R1
50 kΩ
−
µA725
2
3
6
5
R3
5 kΩ
39 Ω
5 kΩ
INPUTS
0.02 µF
R4
5 kΩ
10 Ω
0.05 µF
R7
50 kΩ
+
R6
50 kΩ
OUTPUT
µA725
39 Ω
5 kΩ
0.02 µF
10 Ω
0.05 µF

Pin numbers are shown for metal package only.

Fig. 3-35

WIDE BANDWIDTH, LOW NOISE, LOW DRIFT AMPLIFIER


C2
R2
V_CC
f_max ≥ 240 kHz
10 V
−10 V
R1
V_in
C1
LF156
2
3
7
8
6
4
V_EE

- Power BW: $f_{max} = \frac{S_r}{2\pi V_p} \geq 240$ kHz
- Parasitic input capacitance (C1 ≥ 3 pF for LF155, LF156, and LF157 plus any additional layout capacitance) interacts with feedback elements and creates undesirable high frequency pole To compensate add C2 such that: R2C2 ≥ R1C1.

Fig. 3-36

SIGNAL DISTRIBUTION AMPLIFIER

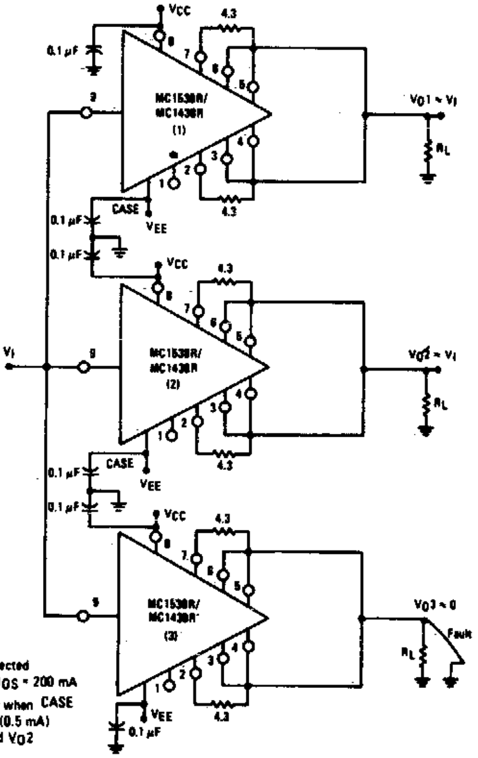

VCC
4.3
0.1 μF
8
7
6
5
9
MC1538R/
MC1438R
(1)
4
3
2
1
VO1 ≈ Vi
RL
CASE
VEE
0.1 μF
0.1 μF
Vi
MC1538R/
MC1438R
(2)
VO2 ≈ Vi
MC1538R/
MC1438R
(3)
VO3 ≈ 0
Fault
0.1 μF

Fig. 3-37

Characteristics

I. Power supplies are protected from the output fault; I_{OS} = 200 mA

II. The small input current when in output current limit (0.5 mA) will isolate V_i, V_O1 and V_O2 from the fault at V_O3.

AUDIO DISTRIBUTION AMPLIFIER

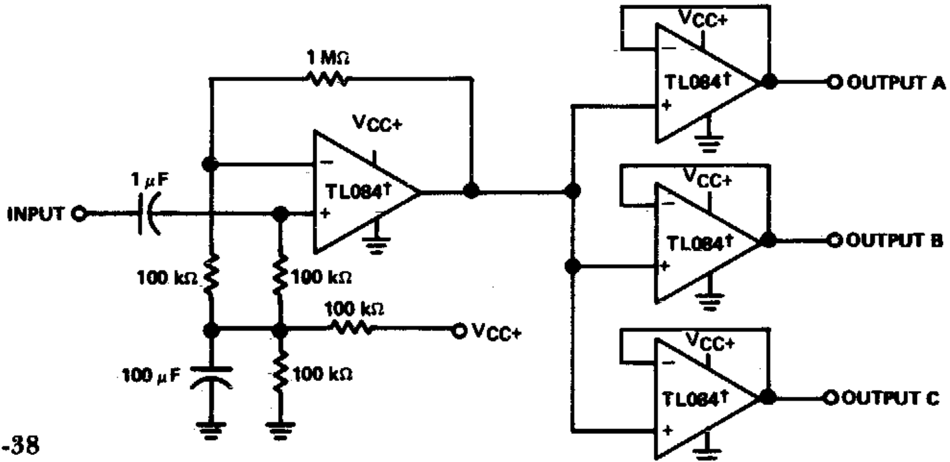

1 MΩ
VCC+
TL084†
1 μF
INPUT
100 kΩ
100 kΩ
100 kΩ
VCC+
100 μF
100 kΩ
VCC+
TL084†
OUTPUT A
VCC+
TL084†
OUTPUT B
VCC+
TL084†
OUTPUT C

Fig. 3-38

HIGH INPUT IMPEDANCE, HIGH OUTPUT CURRENT VOLTAGE FOLLOWER

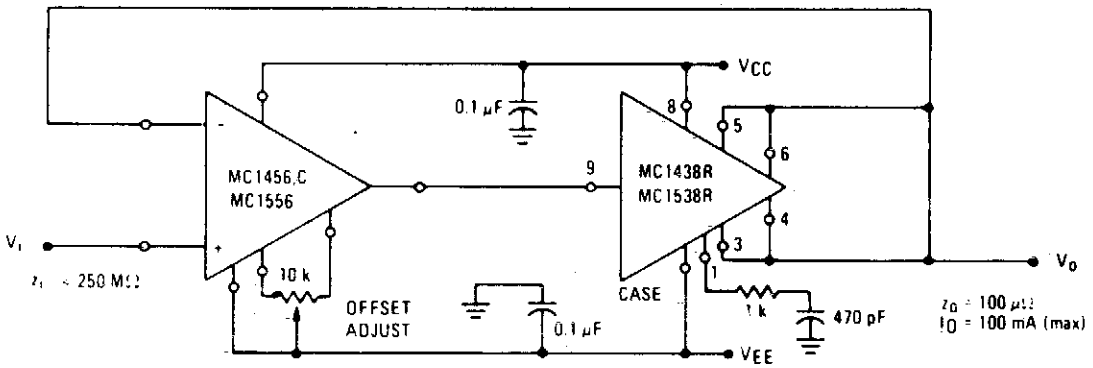
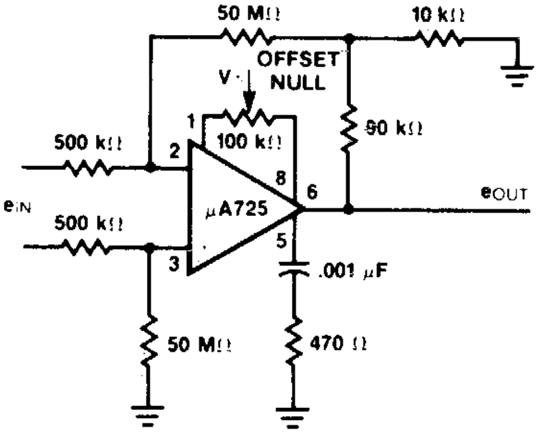

VCC
0.1 μF
8
5
6
9
MC1456,C
MC1556
MC1438R
MC1538R
-
+
4
3
1
Vi
zi ≈ 250 MΩ
10 k
OFFSET
ADJUST
0.1 μF
CASE
1 k
470 pF
VEE
Vo
zo ≈ 100 μΩ
IO = 100 mA (max)

Fig. 3-39

PRECISION AMPLIFIER


50 MΩ
10 kΩ
OFFSET
NULL
V-
1
100 kΩ
90 kΩ
500 kΩ
2
8
6
eOUT
eIN
μA725
500 kΩ
5
3
.001 μF
50 MΩ
470 Ω

Pin numbers are shown for metal package only.

Characteristics

A_V = 1000 = 60 dB

DC Gain Error = 0.05%

Bandwidth = 1 kHz for −0.05% error

Diff. Input Res. = 1 MΩ

Typical amplifying capability

e_{IN} = 10 μV on V_{CMI} = 1.0 V

Caution: Minimize Stray Capacitance

A_{VCL} = 1000

Fig. 3-40

PREAMPLIFIER AND HIGH-TO-LOW IMPEDANCE CONVERTER

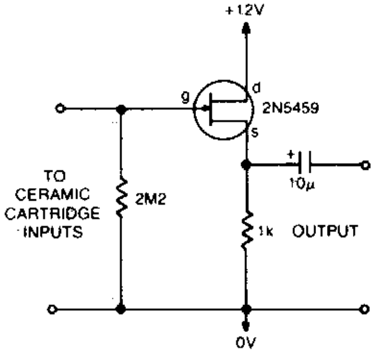

+12V
d
g
2N5459
s
10µ
TO
CERAMIC
CARTRIDGE
INPUTS
2M2
1k
OUTPUT
0V

Fig. 3-41

Circuit Notes

Thiscircuit matches the very high impedance of ceramic cartridges, unity gain, and low impedance output. By "loading" the cartridge with a 2M2 input resistance, the cartridge characteristics are such as to quite closely compensate for the RIAA recording curve. The output from this preamp may be fed to a level pot for mixing.

NONINVERTING AMPLIFIER

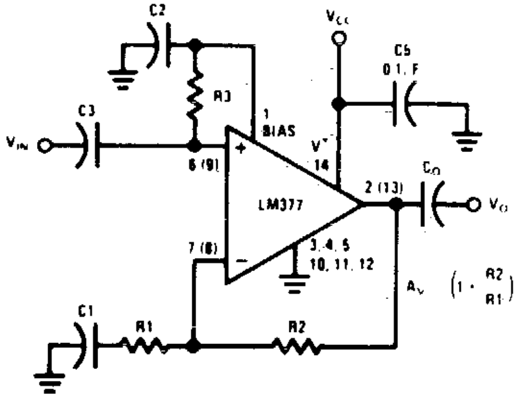
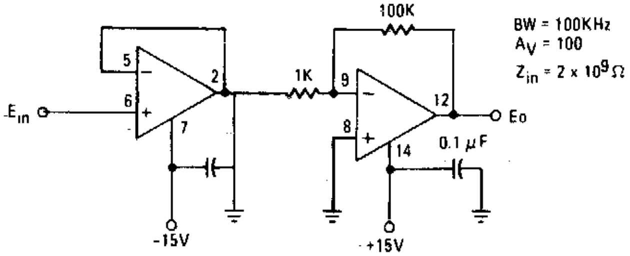

C2
V_{CC}
C5
0.1 F
R3
C3
1
BIAS
V_{IN}
V⁺
14
6 (9)
C_O
2 (13)
LM377
V_O
7 (8)
3, 4, 5
10, 11, 12
A_V (1 + R2/R1)
C1
R1
R2

Fig. 3-42

HIGH IMPEDANCE, HIGH GAIN, HIGH FREQUENCY INVERTING AMP

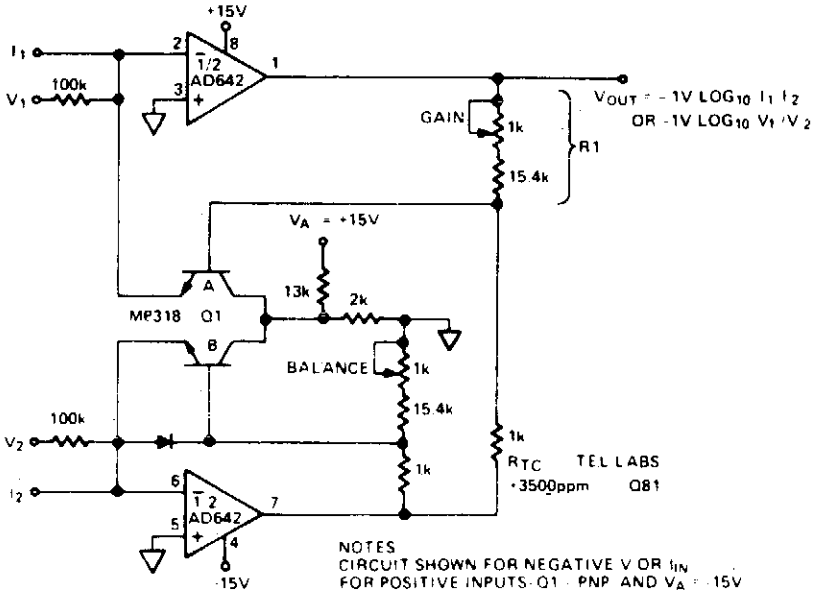

100K
BW = 100KHz
A_V = 100
Z_{in} = 2 x 10⁹Ω
5
2
1K
9
6
12
E_{in}
E_o
7
8
14
0.1 µF
-15V
+15V

IC = HA2650/2655

Fig. 3-43

LOG-RATIO AMPLIFIER

Fig. 3-44


+15V
8
I_1
2
1/2
AD642
1
3
100k
V_1
GAIN
1k
15.4k
R1
$V_{OUT} = -1V \log_{10} I_1/I_2$
OR $-1V \log_{10} V_1/V_2$
$V_A = +15V$
13k
2k
A
MP318 Q1
B
BALANCE
1k
15.4k
100k
V_2
1k
1k
R_{TC} TEL LABS
+3500ppm Q81
I_2
6
1 2
AD642
5
7
4
-15V

NOTES
CIRCUIT SHOWN FOR NEGATIVE V OR I_{IN}
FOR POSITIVE INPUTS Q1 - PNP AND V_A = -15V

INVERTING AMPLIFIER

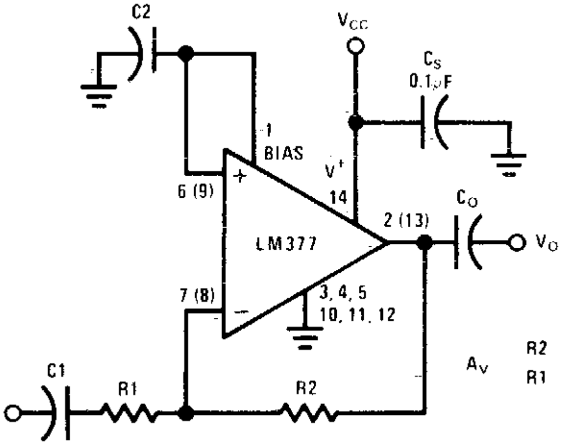

C2
V_{CC}
C_S
0.1µF
1
BIAS
V^+
6 (9)
14
C_O
2 (13)
LM377
V_O
7 (8)
3, 4, 5
10, 11, 12
C1
R1
R2
$A_V = \frac{R2}{R1}$

Fig. 3-45

4

Analog-to-Digital Converters

The sources of the following circuits are contained in the Sources section beginning on page 730. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

8-Bit A/D Converter
Successive Approximation A/D Converter
8-Bit A/D Converter
8-Bit Tracking A/D Converter
8-Bit Successive Approximation A/D Converter
Four Channel Digitally Multiplexed Ramp A/D Converter
Three Decade Logarithmic A/D Converter
Tracking (Servo Type) A/D Converter
3½ Digit A/D Converter with LCD Display
Fast Precision A/D Converter
High Speed 3-Bit A/D Converter
Three IC Low Cost A/D Converter

8-BIT A/D CONVERTER

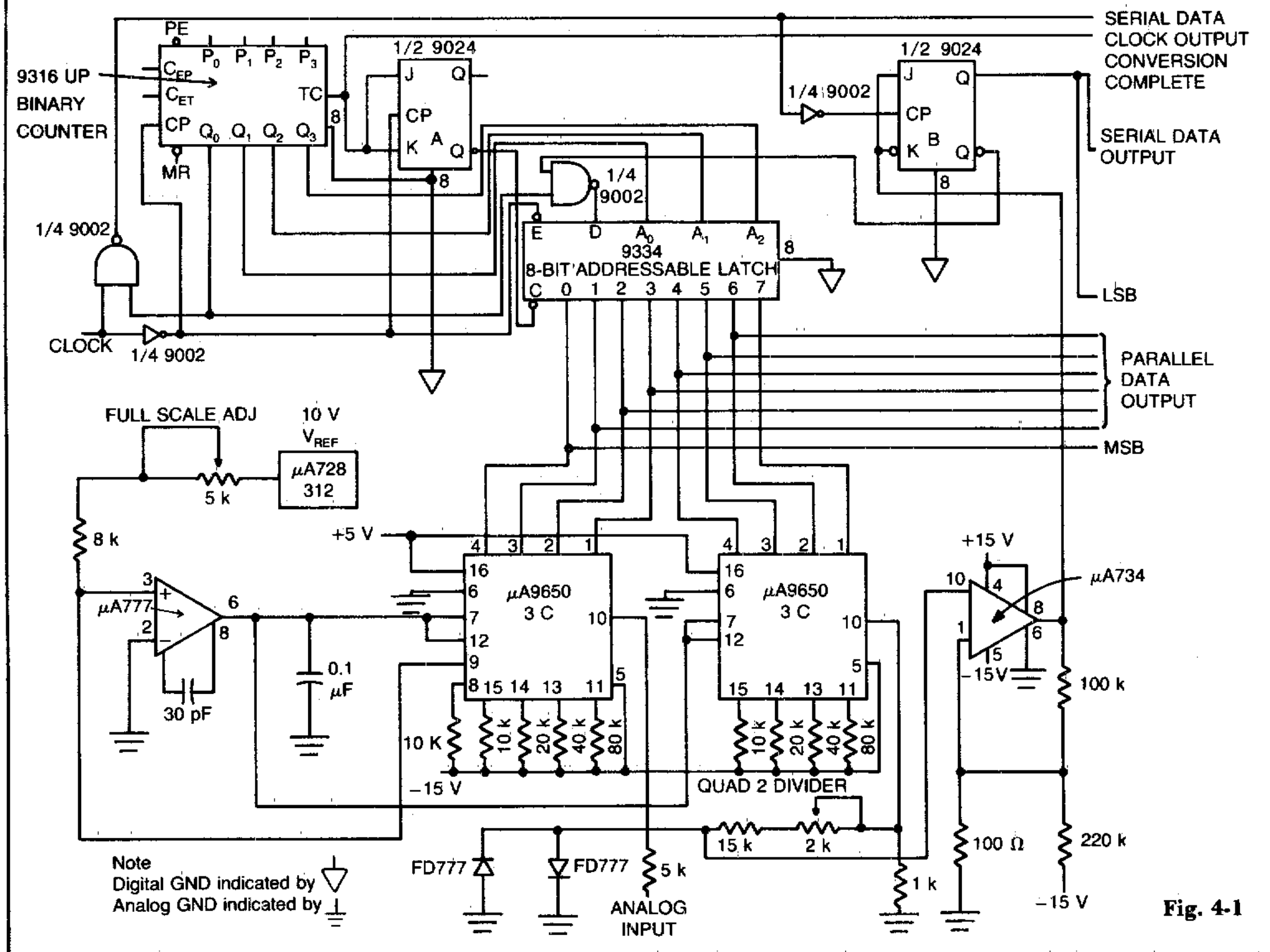

9316 UP
BINARY
COUNTER
PE
C_{EP}
C_{ET}
CP
P_0 P_1 P_2 P_3
TC
Q_0 Q_1 Q_2 Q_3
MR
8
1/4 9002
CLOCK
1/4 9002
1/2 9024
J Q CP K A Q
8
1/4 9002
E D A_0 A_1 A_2
9334
8-BIT ADDRESSABLE LATCH
C 0 1 2 3 4 5 6 7
8
1/4 9002
1/2 9024
J Q CP K B Q
8
SERIAL DATA
CLOCK OUTPUT
CONVERSION
COMPLETE
SERIAL DATA
OUTPUT
LSB
PARALLEL
DATA
OUTPUT
MSB
FULL SCALE ADJ
5 k
10 V
V_{REF}
μA728
312
8 k
3
2
6
8
μA777
30 pF
0.1
μF
+5 V
μA9650
3 C
16 6 7 12 9 8 15 14 13 11 5 10 4 3 2 1
10 K
10 k
20 k
40 k
80 k
–15 V
μA9650
3 C
QUAD 2 DIVIDER
15 k
2 k
FD777
FD777
5 k
ANALOG
INPUT
1 k
+15 V
μA734
10 4 8 6 1 5
–15V
100 k
100 Ω
220 k
–15 V
Note
Digital GND indicated by
Analog GND indicated by

Fig. 4-1

SUCCESSIVE APPROXIMATION A/D CONVERTER

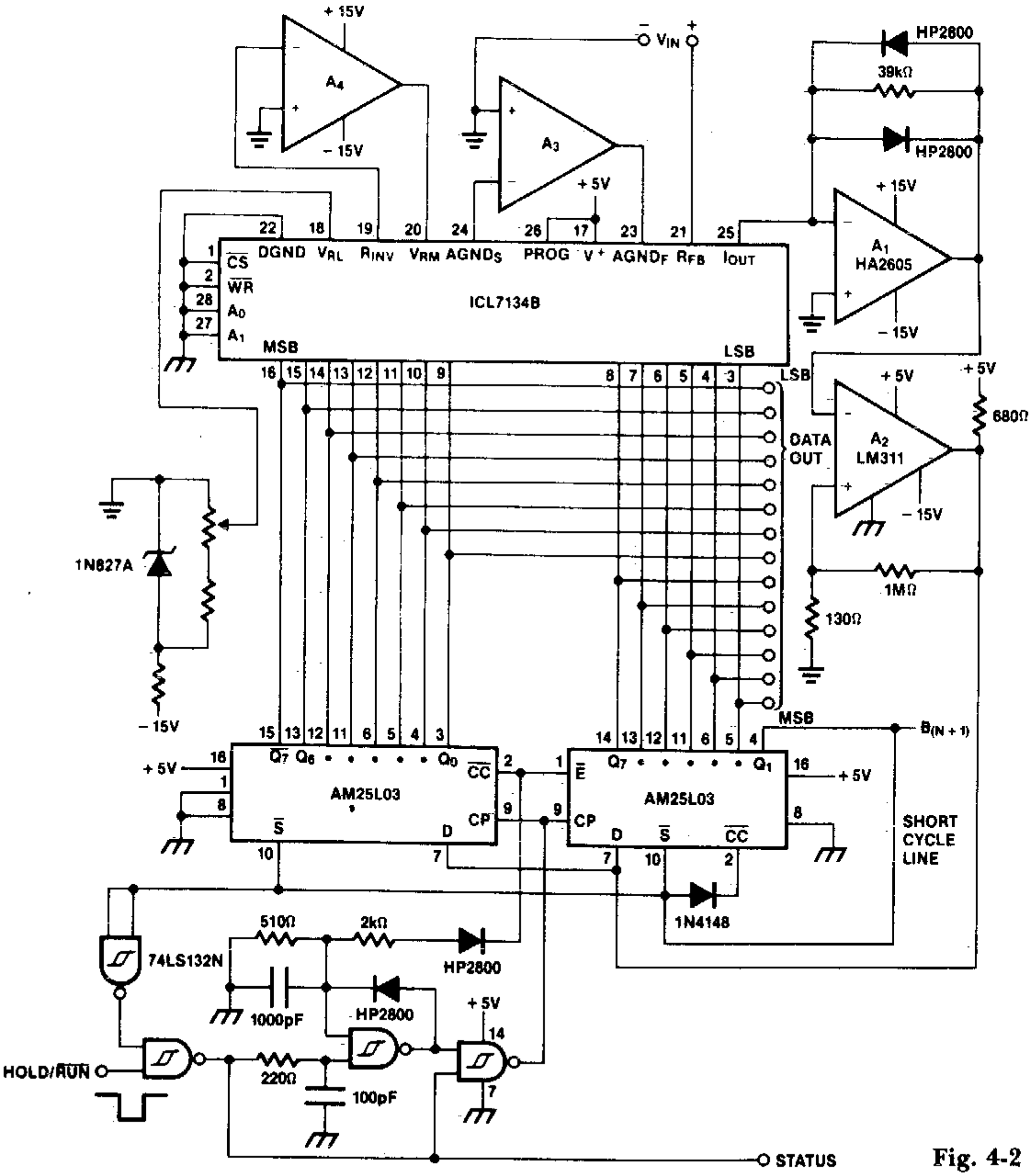

+15V
A4
−15V
A3
+5V
VIN
HP2800
39kΩ
HP2800
+15V
A1
HA2605
−15V
22 DGND
18 VRL
19 RINV
20 VRM
24 AGNDS
26 PROG
17 V+
23 AGNDF
21 RFB
25 IOUT
1 CS
2 WR
28 A0
27 A1
ICL7134B
MSB
LSB
16 15 14 13 12 11 10 9
8 7 6 5 4 3
LSB
+5V
+5V
680Ω
DATA OUT
A2
LM311
−15V
1MΩ
130Ω
1N827A
−15V
MSB
B(N+1)
+5V
15 13 12 11 6 5 4 3
16 1 8
Q7 Q6 Q0
CC
AM25L03
S
D
CP
2
9
10
7
1 E
14 13 12 11 6 5 4
Q7 Q1
16
+5V
AM25L03
9 CP
D
S
CC
8
7 10 2
SHORT CYCLE LINE
1N4148
74LS132N
510Ω
2kΩ
HP2800
1000pF
HP2800
+5V
14
7
HOLD/RUN
220Ω
100pF
STATUS

Fig. 4-2

Circuit Notes

A bipolar input, high speed A/D converter uses two AM25L03s to form a 14-bit successive approximation register. The comparator is a two-stage circuit with an HA2605 front-end amplifier used to reduce settling time problems at the summing node. Careful offset-nulling of this amplifier is needed.

8-BIT A/D CONVERTER

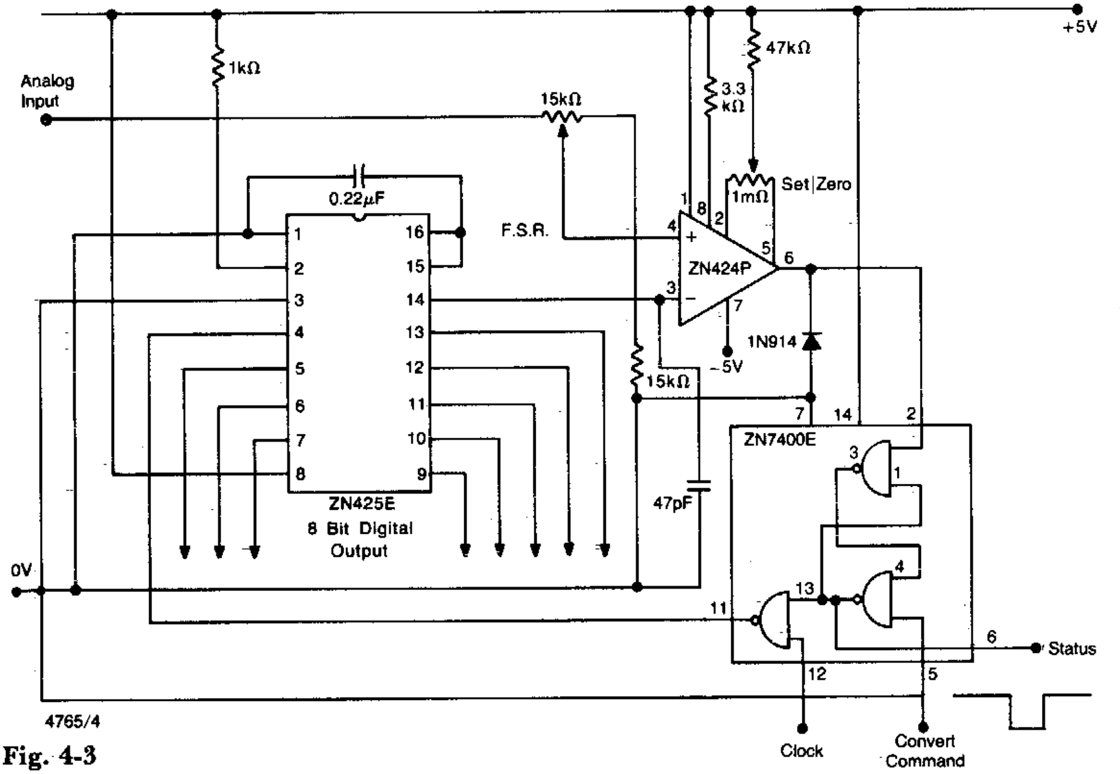

+5V
47kΩ
1kΩ
3.3 kΩ
Analog Input
15kΩ
0.22μF
1mΩ
Set|Zero
F.S.R.
ZN424P
1N914
−5V
15kΩ
ZN425E
8 Bit Digital Output
47pF
ZN7400E
0V
Status
4765/4
Clock
Convert Command

Fig. 4-3

8-BIT TRACKING A/D CONVERTER

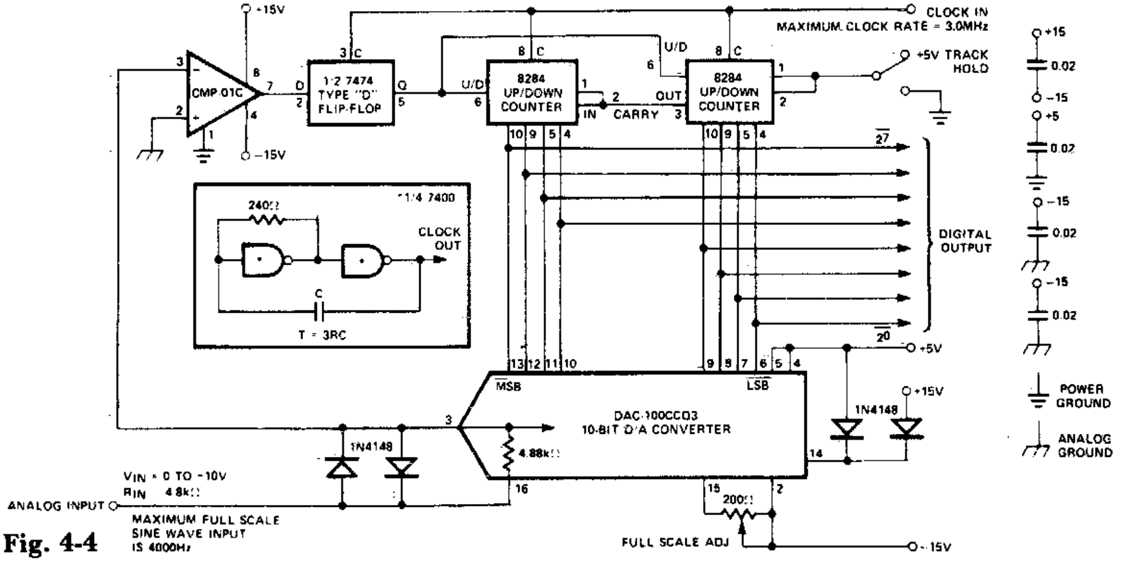

+15V
−15V
CMP-01C
1/2 7474 TYPE "D" FLIP-FLOP
8284 UP/DOWN COUNTER
8284 UP/DOWN COUNTER
U/D
CARRY
OUT
IN
CLOCK IN
MAXIMUM CLOCK RATE = 3.0MHz
+5V TRACK
HOLD
1/4 7400
240Ω
CLOCK OUT
C
T = 3RC
DIGITAL OUTPUT
MSB
LSB
DAC-100CC03
10-BIT D/A CONVERTER
4.88kΩ
1N4148
1N4148
+5V
+15V
200Ω
FULL SCALE ADJ
−15V
VIN = 0 TO −10V
RIN 4.8kΩ
ANALOG INPUT
MAXIMUM FULL SCALE SINE WAVE INPUT IS 4000Hz
+15
0.02
−15
+5
0.02
−15
0.02
−15
0.02
POWER GROUND
ANALOG GROUND

Fig. 4-4

8-BIT SUCCESSIVE APPROXIMATION A/D CONVERTER

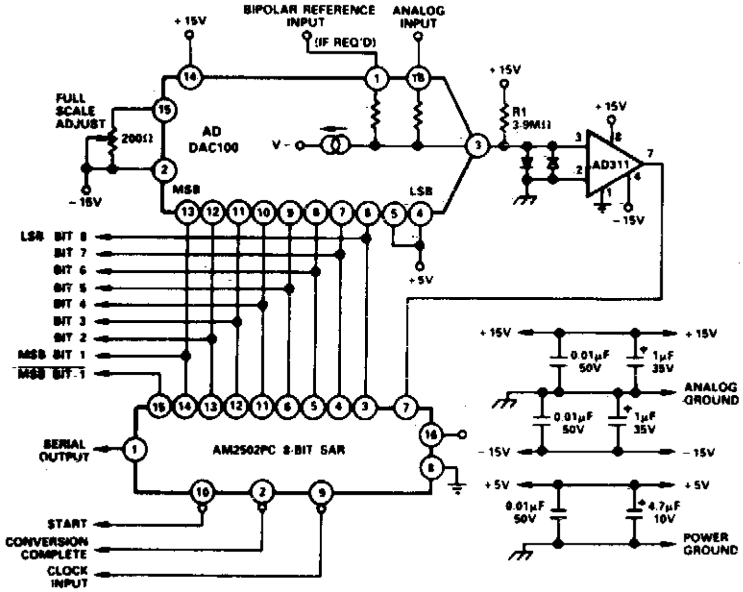

+15V
BIPOLAR REFERENCE INPUT
(IF REQ'D)
ANALOG INPUT
FULL SCALE ADJUST
200Ω
−15V
AD DAC100
V−
MSB
LSB
+15V
R1 3.9MΩ
+15V
AD311
−15V
+5V
LSB BIT 8
BIT 7
BIT 6
BIT 5
BIT 4
BIT 3
BIT 2
MSB BIT 1
MSB BIT 1
SERIAL OUTPUT
AM2502PC 8-BIT SAR
START
CONVERSION COMPLETE
CLOCK INPUT
+15V
0.01μF 50V
1μF 35V
ANALOG GROUND
0.01μF 50V
1μF 35V
−15V
+5V
0.01μF 50V
4.7μF 10V
POWER GROUND

Fig. 4-5

FOUR CHANNEL DIGITALLY MULTIPLEXED RAMP A/D CONVERTER

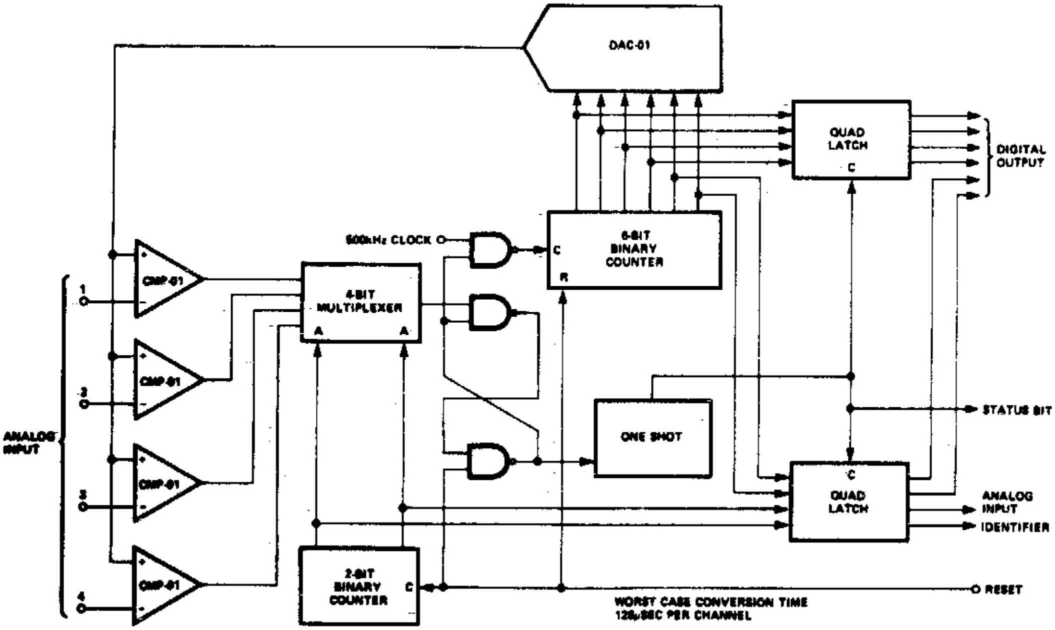

DAC-01
QUAD LATCH
C
DIGITAL OUTPUT
500kHz CLOCK
6-BIT BINARY COUNTER
C
R
CMP-01
4-BIT MULTIPLEXER
A
A
ANALOG INPUT
1
2
3
4
ONE SHOT
STATUS BIT
QUAD LATCH
C
ANALOG INPUT IDENTIFIER
2-BIT BINARY COUNTER
C
RESET
WORST CASE CONVERSION TIME 128μSEC PER CHANNEL

Fig. 4-6

THREE-DECADE LOGARITHMIC A/D CONVERTER

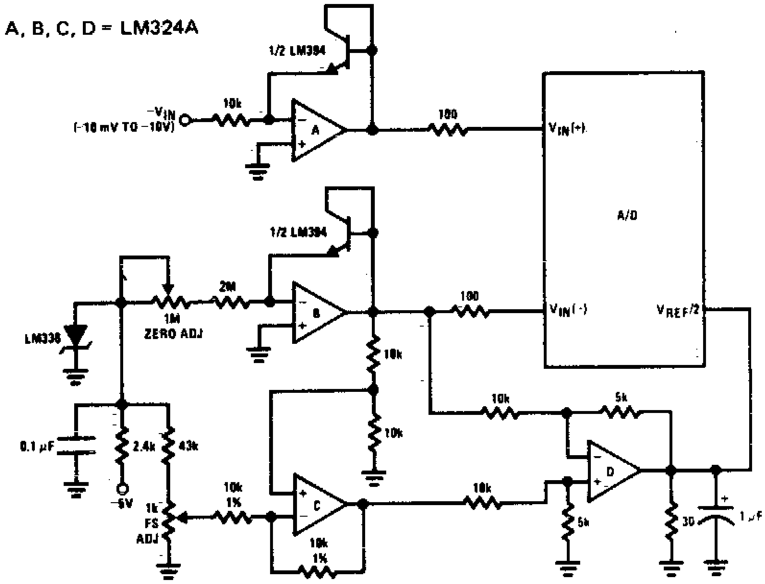

A, B, C, D = LM324A
1/2 LM394
−VIN
(−10 mV TO −10V)
10k
A
100
VIN(+)
A/D
1/2 LM394
2M
1M
ZERO ADJ
LM336
B
100
VIN(−)
VREF/2
10k
10k
10k
5k
0.1 µF
2.4k
43k
−5V
1k
FS
ADJ
10k
1%
C
10k
1%
10k
D
5k
30
1 µF

Fig. 4-7

TRACKING (SERVO TYPE) A/D CONVERTER

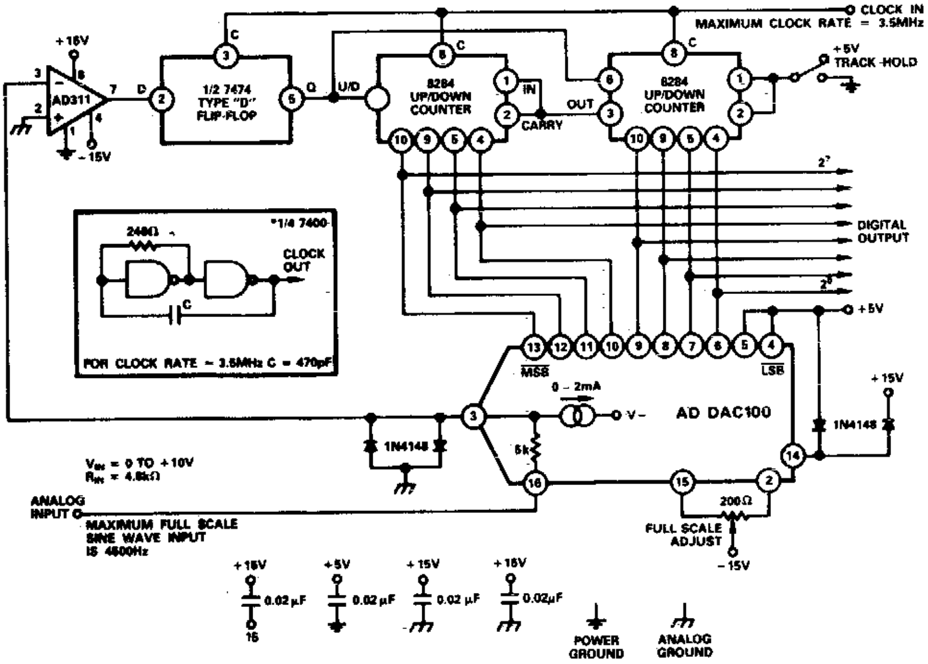
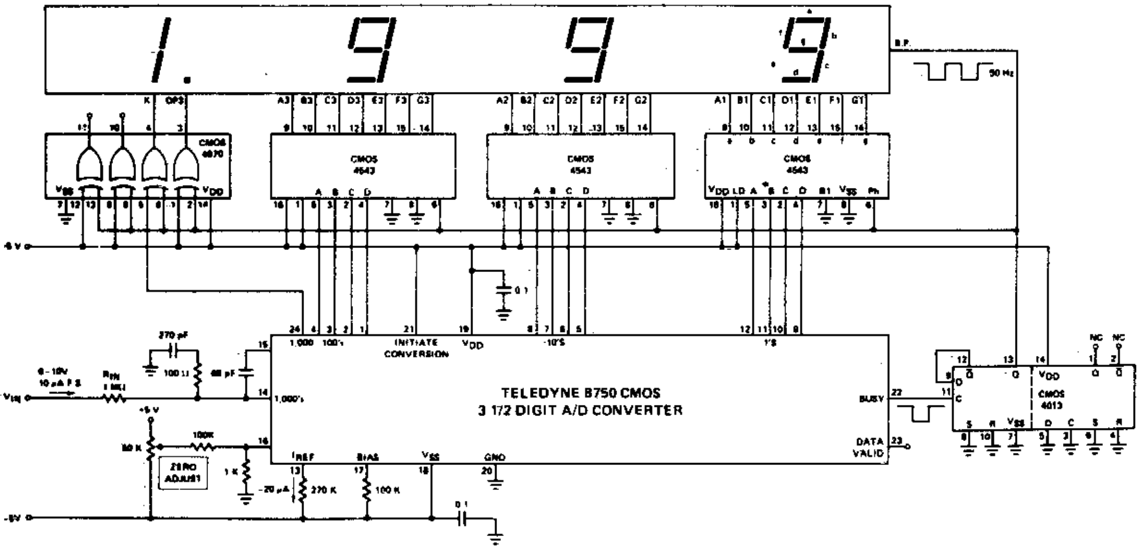

CLOCK IN
MAXIMUM CLOCK RATE = 3.5MHz
+15V
AD311
−15V
D
1/2 7474
TYPE "D"
FLIP-FLOP
C
Q
U/D
8284
UP/DOWN
COUNTER
IN
CARRY
OUT
8284
UP/DOWN
COUNTER
+5V
TRACK-HOLD
DIGITAL
OUTPUT
*1/4 7400
240Ω
CLOCK
OUT
C
FOR CLOCK RATE ~ 3.5MHz C = 470pF
+5V
MSB
LSB
0 – 2mA
V−
AD DAC100
+15V
1N4148
1N4148
5k
VIN = 0 TO +10V
RIN = 4.8kΩ
ANALOG
INPUT
MAXIMUM FULL SCALE
SINE WAVE INPUT
IS 4600Hz
200Ω
FULL SCALE
ADJUST
−15V
+15V
0.02 µF
+5V
0.02 µF
+15V
0.02 µF
+15V
0.02µF
POWER
GROUND
ANALOG
GROUND

Fig. 4-8

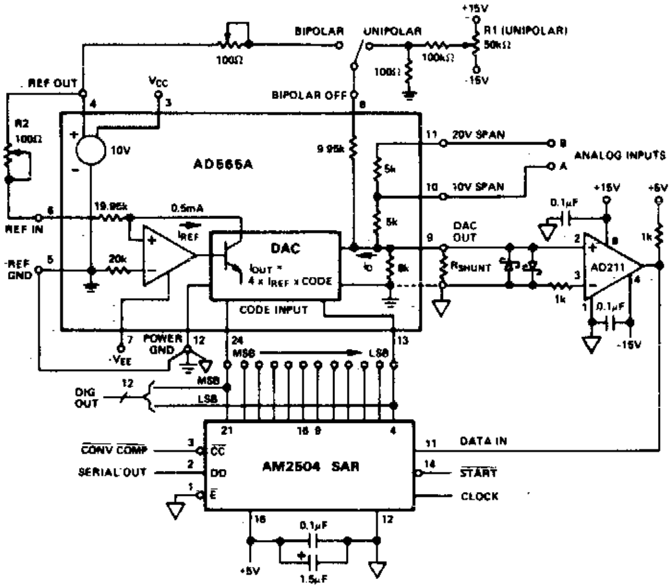
3½ DIGIT A/D CONVERTER WITH LCD DISPLAY



1. 9 9 9
B.P.
50 Hz
CMOS 4070
CMOS 4543
CMOS 4543
CMOS 4543
+5 V
0.1
270 pF
0 - 10V 10 μA F.S.
R_{IN} 1 MΩ
V_{IN}
100 Ω
68 pF
50 K
100K
ZERO ADJUST
1 K
-20 μA
270 K
100 K
-5V
0.1
INITIATE CONVERSION
V_{DD}
TELEDYNE 8750 CMOS
3 1/2 DIGIT A/D CONVERTER
BUSY
DATA VALID
I_{REF}
BIAS
V_{SS}
GND
1,000
100's
10's
1's
1,000's
CMOS 4013
NC
NC

Fig. 4-9

FAST PRECISION A/D CONVERTER



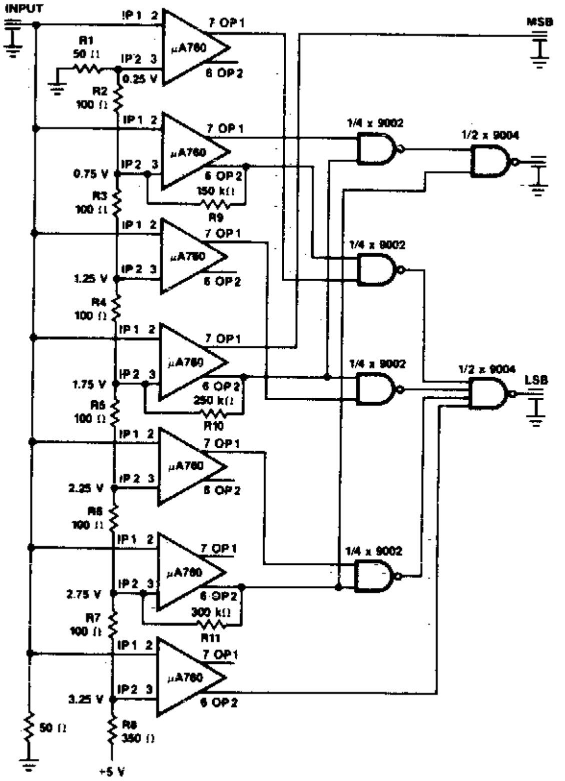
+15V
BIPOLAR
UNIPOLAR
R1 (UNIPOLAR) 50kΩ
-15V
100Ω
100kΩ
100Ω
REF OUT
V_{CC}
BIPOLAR OFF
R2 100Ω
10V
AD565A
9.95k
5k
5k
8k
20V SPAN
10V SPAN
ANALOG INPUTS
B
A
REF IN
19.95k
0.5mA
I_{REF}
20k
-REF GND
DAC
I_{OUT} = 4 x I_{REF} x CODE
I_O
DAC OUT
R_{SHUNT}
0.1μF
+15V
+5V
1k
AD211
1k
0.1μF
-15V
CODE INPUT
POWER GND
$-V_{EE}$
MSB
LSB
DIG OUT
AM2504 SAR
$\overline{CONV\ COMP}$
SERIAL OUT
$\overline{CC}$
DO
$\overline{E}$
DATA IN
$\overline{START}$
CLOCK
0.1μF
1.5μF
+5V

INPUT RANGES

Unipolar	Bipolar	Connect	Equiv DAC Z_{OUT}
0 TO +10	±5	INPUT TO A	2.35kΩ
0 TO +5	±2.5	INPUT TO A	1.90kΩ
0 TO +20	±10	INPUT TO B B TO DAC OUT	3.08kΩ

Fig. 4-10

HIGH SPEED 3-BIT A/D CONVERTER


INPUT
R1
50 Ω
R2
100 Ω
R3
100 Ω
R4
100 Ω
R5
100 Ω
R6
100 Ω
R7
100 Ω
R8
350 Ω
50 Ω
+5 V
0.25 V
0.75 V
1.25 V
1.75 V
2.25 V
2.75 V
3.25 V
IP1 2
IP2 3
7 OP1
6 OP2
μA760
150 kΩ
R9
250 kΩ
R10
300 kΩ
R11
1/4 x 9002
1/2 x 9004
MSB
LSB

Input voltage range: 3.5 V

Typical conversion speed: 30 ns

Fig. 4-11

THREE IC LOW COST A/D CONVERTER

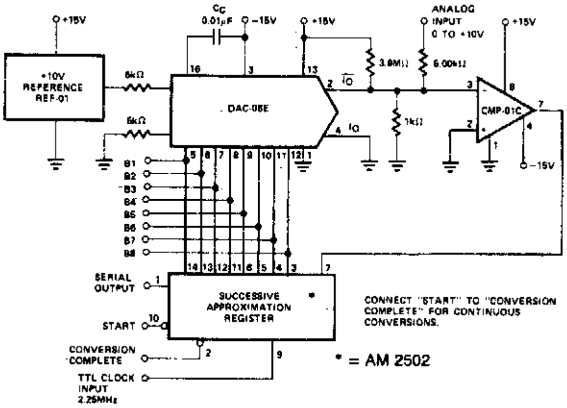

+15V
+10V REFERENCE REF-01
6kΩ
6kΩ
C_C 0.01μF
−15V
+15V
DAC-08E
$\overline{I_O}$
I_O
3.9MΩ
5.00kΩ
1kΩ
ANALOG INPUT 0 TO +10V
+15V
CMP-01C
−15V
B1
B2
B3
B4
B5
B6
B7
B8
SERIAL OUTPUT
START
CONVERSION COMPLETE
TTL CLOCK INPUT 2.25MHz
SUCCESSIVE APPROXIMATION REGISTER *
CONNECT "START" TO "CONVERSION COMPLETE" FOR CONTINUOUS CONVERSIONS.
* = AM 2502

Fig. 4-12

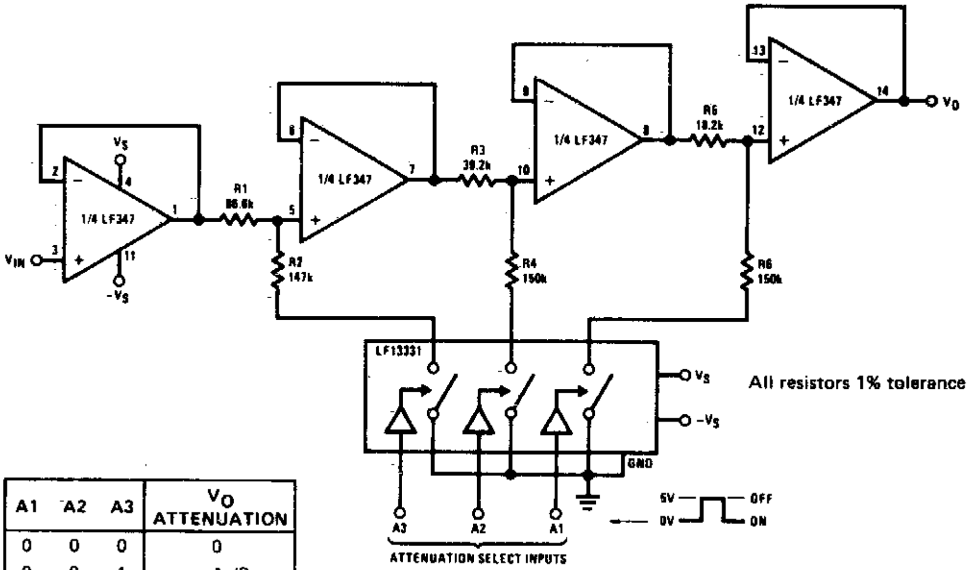
5

Attenuators

The sources of the following circuits are contained in the Sources section beginning on page 730. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

Digitally Selectable Precision Attenuator
Variable Attenuator
Digitally Controlled Amplifier/Attenuator
Programmable Attenuator (1 to 0.0001)

DIGITALLY SELECTABLE PRECISION ATTENUATOR


1/4 LF347
1/4 LF347
1/4 LF347
1/4 LF347
V_S
$-V_S$
V_{IN}
V_O
R1 86.6k
R2 147k
R3 39.2k
R4 150k
R5 18.2k
R6 150k
1
2
3
4
5
6
7
8
9
10
11
12
13
14
LF13331
V_S
$-V_S$
GND
A3
A2
A1
ATTENUATION SELECT INPUTS
5V
0V
OFF
ON

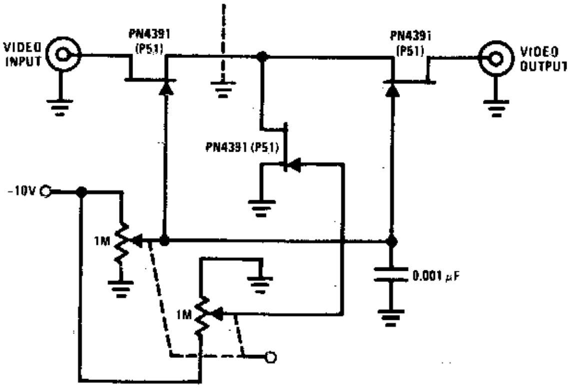
All resistors 1% tolerance

Fig. 5-1

A1	A2	A3	V_O ATTENUATION
0	0	0	0
0	0	1	−1 dB
0	1	0	−2 dB
0	1	1	−3 dB
1	0	0	−4 dB
1	0	1	−5 dB
1	1	0	−6 dB
1	1	1	−7 dB

- Accuracy of better than 0.4% with standard 1% value resistors
- No offset adjustment necessary
- Expandable to any number of stages
- Very high input impedance

VARIABLE ATTENUATOR

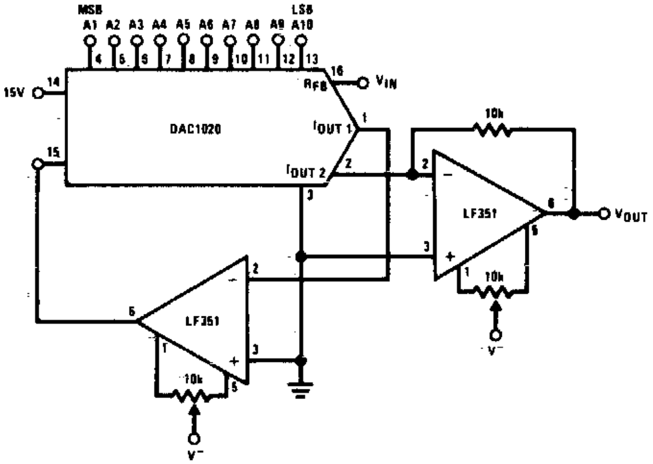

VIDEO INPUT
PN4391 (P51)
PN4391 (P51)
PN4391 (P51)
VIDEO OUTPUT
-10V
1M
1M
0.001 µF

Circuit Notes

The PN4391 provides a low $R_{ds(on)}$ (less than 30 ohms). The tee attenuator provides for optimum dynamic linear range for attenuation and if complete turn-off is desired, attenuation of greater than 100 dB can be obtained at 10 MHz providing proper rf construction techniques are employed.

Fig. 5-2

DIGITALLY CONTROLLED AMPLIFIER/ATTENUATOR


MSB
LSB
A1 A2 A3 A4 A5 A6 A7 A8 A9 A10
4 5 6 7 8 9 10 11 12 13
15V
14
15
16
DAC1020
R_{FB}
V_{IN}
$I_{OUT\,1}$
$I_{OUT\,2}$
1
2
3
10k
LF351
6
5
V_{OUT}
V⁻

$$V_{OUT} = V_{REF}\left[\frac{\dfrac{\overline{A1}}{2} + \dfrac{\overline{A2}}{4} + \cdots + \dfrac{A10}{1024}}{\dfrac{A1}{2} + \dfrac{A2}{4} + \cdots + \dfrac{\overline{A10}}{1024}}\right] \text{ or } V_{OUT} = V_{REF}\left(\frac{1023 - N}{N}\right)$$

where: $0 \leq N \leq 1023$

$N = 0$ for A_N = all zeros

$N = 1$ for $A10 = 1$, $A1{-}A9 = 0$

.

.

.

$N = 1023$ for A_N = all 1's

Fig. 5-3

PROGRAMMABLE ATTENUATOR (1 TO 0.0001)

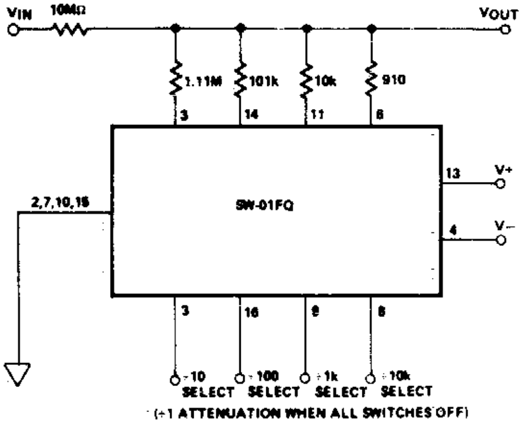

V_{IN}
10MΩ
V_{OUT}
1.11M
101k
10k
910
3
14
11
6
SW-01FQ
2,7,10,15
13
V+
4
V−
3
16
9
8
÷10 SELECT
÷100 SELECT
÷1k SELECT
÷10k SELECT
(÷1 ATTENUATION WHEN ALL SWITCHES OFF)

Fig. 5-4

6

Audio Mixers

The sources of the following circuits are contained in the Sources section beginning on page 730. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

Four Input Stereo Mixer
High-Level Four-Channel Mixer
Two Channel Panning Circuit
CMOS Mixer
Mixer Preamplifier with Tone Control
Passive Mixer
One Transistor Audio Mixer
Silent Audio Switching/Mixing
Hybrid Mixer
Four Channel Mixer

FOUR-INPUT STEREO MIXER

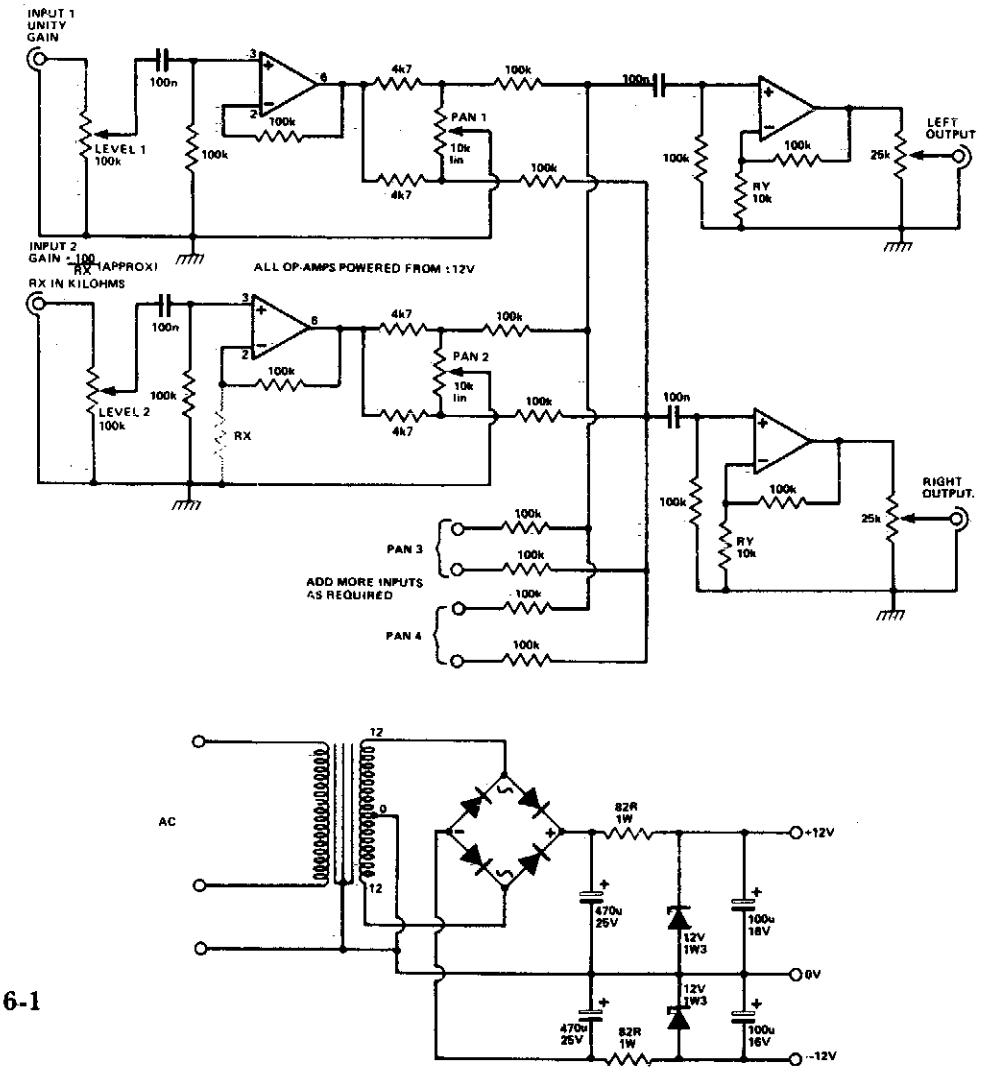
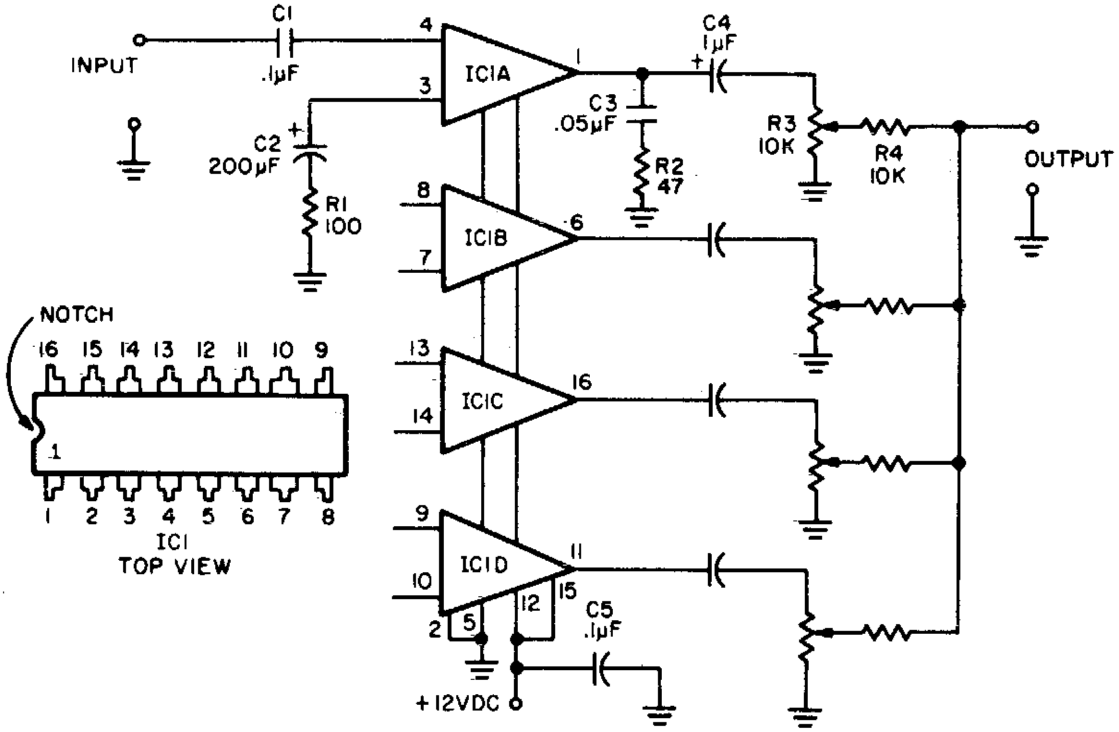

INPUT 1
UNITY
GAIN
100n
100k
LEVEL 1
100k
3
2
6
4k7
100k
PAN 1
10k
lin
4k7
100k
100n
100k
RY
10k
25k
LEFT
OUTPUT
INPUT 2
GAIN = 100/RX (APPROX)
RX IN KILOHMS
ALL OP-AMPS POWERED FROM ±12V
LEVEL 2
100k
RX
PAN 2
RIGHT
OUTPUT.
PAN 3
ADD MORE INPUTS
AS REQUIRED
PAN 4
AC
12
0
12
82R
1W
+12V
470u
25V
12V
1W3
100u
16V
0V
-12V

Fig. 6-1

Circuit Notes

Four (or more) inputs can be mixed and produce stereo output. Gain of each stage can be boosted by adding RX, but it should be kept below 50 (RX above 2.2 K) to avoid poor frequency, response. If more than four stages are used, decrease RX to 6.8 K for six inputs, or 4.7 K for eight inputs. The op amps are 741 or other lower noise types. The power supply circuit is also given.

HIGH-LEVEL FOUR-CHANNEL MIXER


C1
.1μF
INPUT
4
3
IC1A
1
C4
1μF
+
C3
.05μF
R2
47
R3
10K
R4
10K
OUTPUT
C2
200μF
+
R1
100
8
7
IC1B
6
NOTCH
16 15 14 13 12 11 10 9
1 2 3 4 5 6 7 8
1
IC1
TOP VIEW
13
14
IC1C
16
9
10
IC1D
11
15
12
2
5
C5
.1μF
+12VDC

PARTS LIST FOR HI-LEVEL MIXER

C1—0.1-uF, 3 VDC capacitor
C2—200-uF, 3 VDC capacitor
C3—0.05-uF, 75 VDC disc capacitor
C4—1-uF, 15 VDC capacitor
C5—0.1-uF, 15 VDC capacitor
IC1—RCA CA 3052
R1—100-ohms, ½-watt resistor
R2—47-ohms, ½-watt resistor
R3—Potentiometer, 10,000-ohms audio taper
R4—10,000-ohms, ½-watt resistor

Fig. 6-2

Circuit Notes

To provide good signal-to-noise ratio, this four channel mixer amplifier controls the signal levels after the amplifiers, and then mixes them to offer a combined output. The circuit works with any 50 ohm to 50 K dynamic microphone but not with crystal or ceramic mikes because the IC input impedance is low. Note that all four circuits are identical but that only one is shown complete.

TWO CHANNEL PANNING CIRCUIT

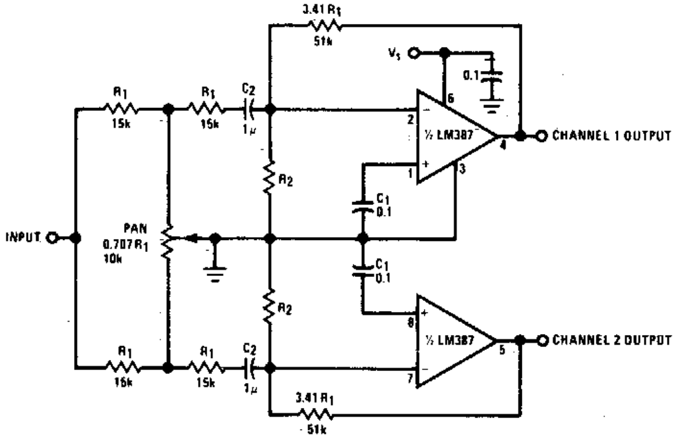

3.41 R1
51k
Vs
0.1
R1
15k
R1
15k
C2
1μ
2
6
½ LM387
4
CHANNEL 1 OUTPUT
+
1
3
R2
C1
0.1
INPUT
PAN
0.707 R1
10k
C1
0.1
R2
8
+
½ LM387
5
CHANNEL 2 OUTPUT
R1
15k
R1
15k
C2
1μ
7
3.41 R1
51k

Fig. 6-3

Circuit Notes

This panning circuit (short for panoramic control circuit) provides the ability to move the apparent position of one microphone's input between two output channels. This effect is often required in recording studio mixing consoles. Panning is how recording engineers manage to pick up your favorite pianist and "float" the sound over to the other side of the stage and back again.

CMOS MIXER

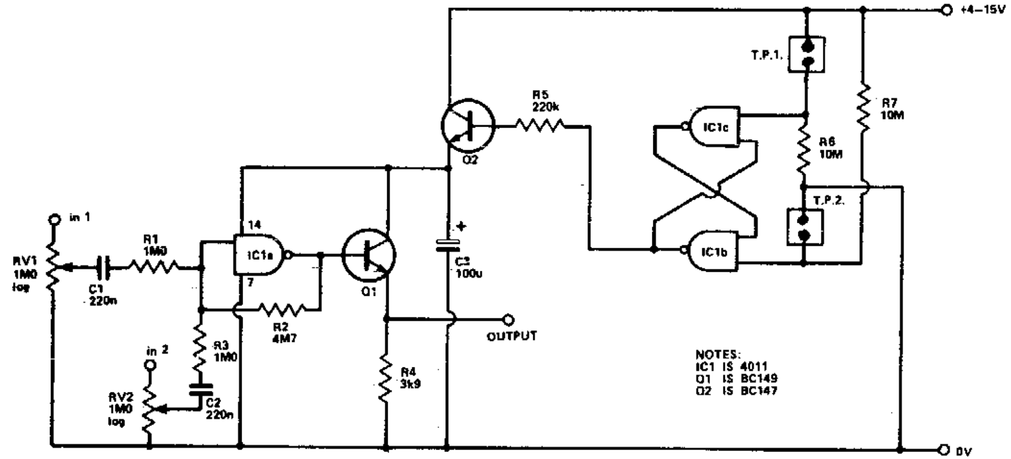

+4–15V
T.P.1.
R5
220k
R7
10M
IC1c
R6
10M
Q2
T.P.2.
in 1
R1
1M0
14
IC1a
RV1
1M0
log
C1
220n
7
Q1
+
C3
100u
IC1b
R2
4M7
OUTPUT
in 2
R3
1M0
R4
3k9
RV2
1M0
log
C2
220n
NOTES:
IC1 IS 4011
Q1 IS BC149
Q2 IS BC147
0V

Fig. 6-4

Circuit Notes

Four inputs can be mixed by duplicating the circuit to the left of C3 and using the fourth gate of IC1. Two gates are used in a touch-operated switching circuit that controls the voltage on the base of switching transistor Q2. Touching TP1 and TP2 alternately turns the circuit on and off.

MIXER PREAMPLIFIER WITH TONE CONTROL

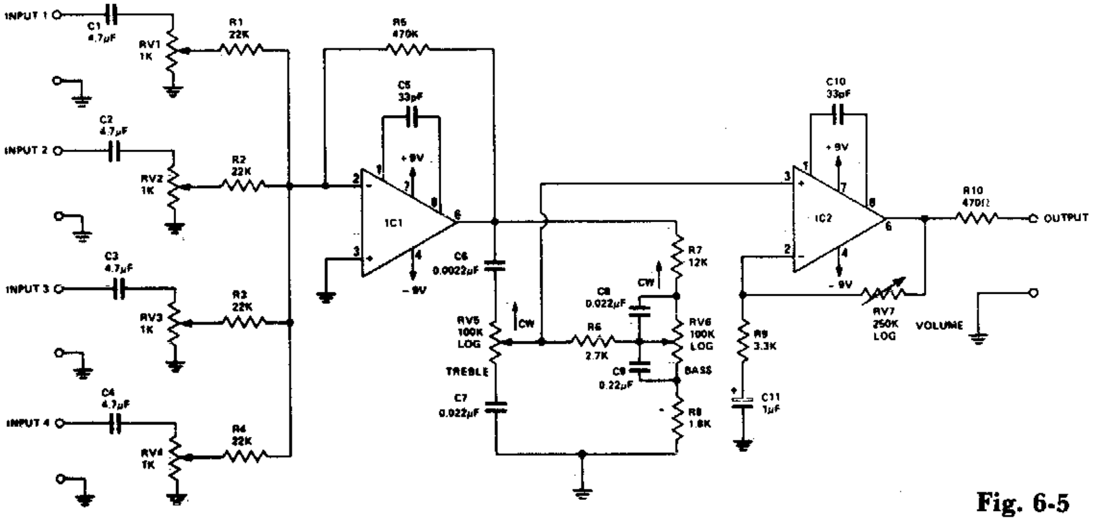

INPUT 1
C1 4.7μF
RV1 1K
R1 22K
INPUT 2
C2 4.7μF
RV2 1K
R2 22K
INPUT 3
C3 4.7μF
RV3 1K
R3 22K
INPUT 4
C4 4.7μF
RV4 1K
R4 22K
R5 470K
C5 33pF
+9V
-9V
IC1
1
2
3
4
6
7
8
C6 0.0022μF
RV5 100K LOG
TREBLE
CW
C7 0.022μF
C8 0.022μF
R6 2.7K
C9 0.22μF
R7 12K
RV6 100K LOG
BASS
R8 1.8K
C10 33pF
IC2
R9 3.3K
C11 1μF
RV7 250K LOG
VOLUME
R10 470Ω
OUTPUT

Fig. 6-5

Circuit Notes

General purpose preamplifier/mixer accepts up to four inputs, has a gain of 1600, and provides bass and treble controls that can be varied ± 10 dB at 100 Hz and 10 kHz respectively. IC1 and IC2 = LM301A.

PASSIVE MIXER

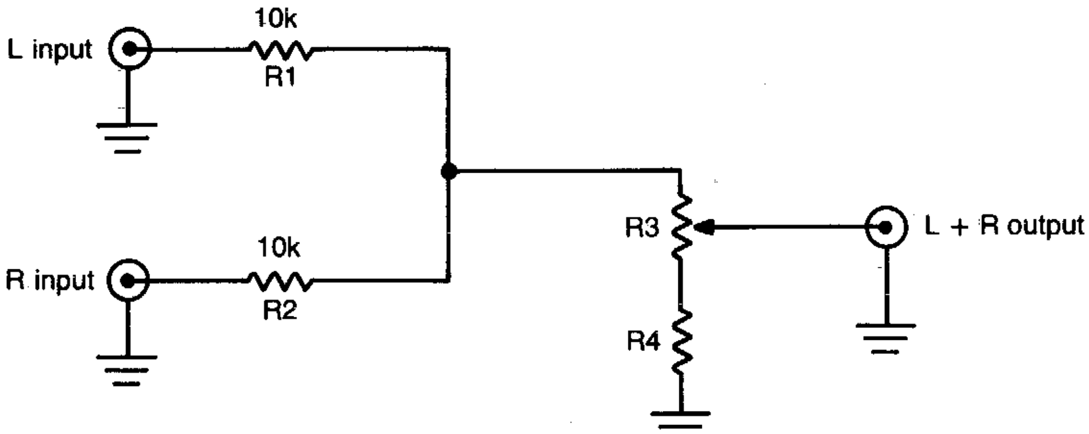
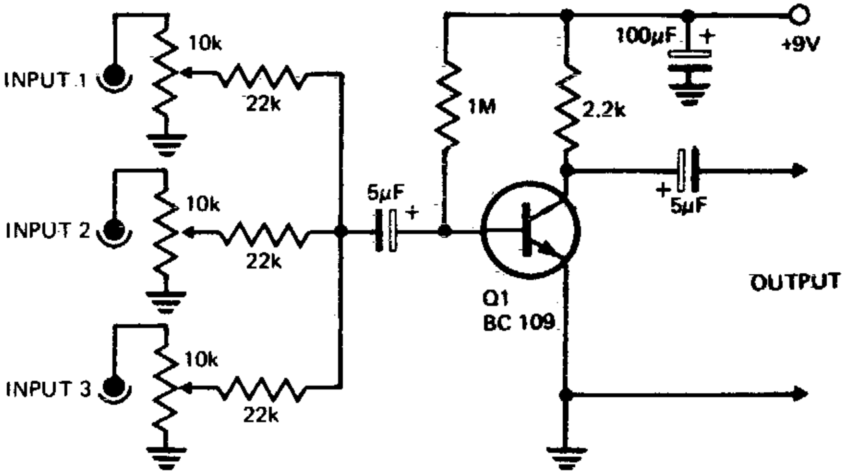

L input
10k
R1
R input
10k
R2
R3
R4
L + R output

Fig. 6-6

Circuit Notes

This simple circuit can be used to combine stereo signals to produce a monaural output. R1 and R2 isolate both circuits and R3 controls the level of the combined output signal.

ONE TRANSISTOR AUDIO MIXER



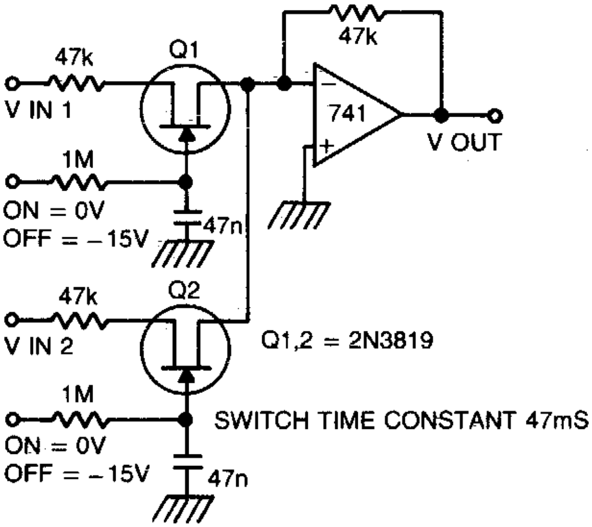
INPUT 1
INPUT 2
INPUT 3
10k
10k
10k
22k
22k
22k
5μF
+
1M
2.2k
100μF
+
+9V
+
5μF
Q1
BC 109
OUTPUT

Fig. 6-7

Circuit Notes

Three or more inputs with individual level controls feed into the base of Q1 that provides a voltage gain of 20.

SILENT AUDIO SWITCHING/MIXING



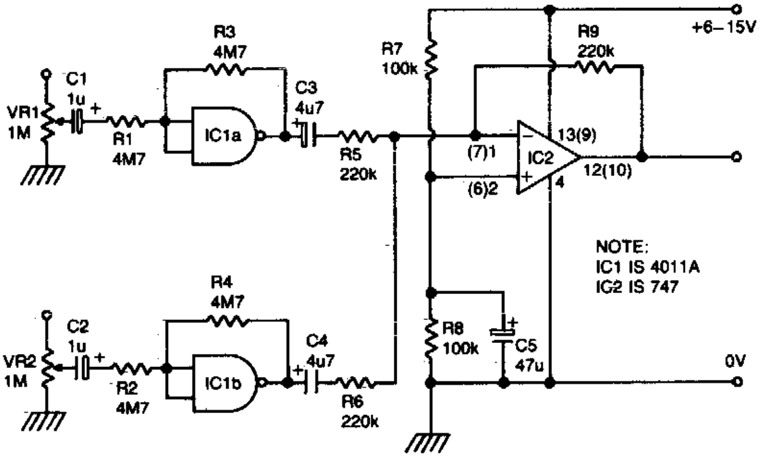
47k
Q1
V IN 1
1M
ON = 0V
OFF = −15V
47n
47k
741
−
+
V OUT
47k
Q2
V IN 2
Q1,2 = 2N3819
1M
SWITCH TIME CONSTANT 47mS
ON = 0V
OFF = −15V
47n

Fig. 6-8

Circuit Notes

Two or more signals can be switched and/or mixed without annoying clicks by using FETs and a low input-impedance op amp circuit.

HYBRID MIXER



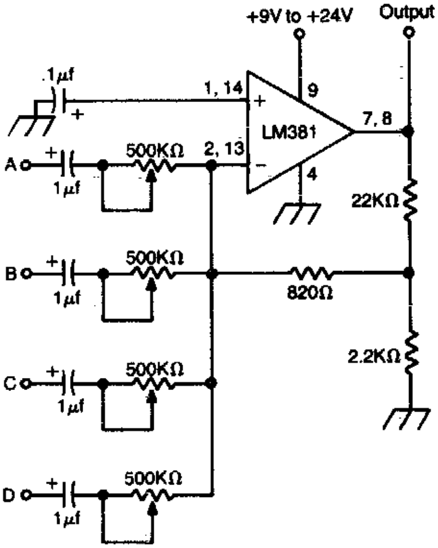
R3
4M7
C1
1u
VR1
1M
R1
4M7
IC1a
C3
4u7
R5
220k
R7
100k
R9
220k
+6–15V
(7)1
(6)2
IC2
13(9)
12(10)
4
NOTE:
IC1 IS 4011A
IC2 IS 747
R4
4M7
C2
1u
VR2
1M
R2
4M7
IC1b
C4
4u7
R6
220k
R8
100k
C5
47u
0V

Fig. 6-9

Circuit Notes

IC1a and b are biased into the linear regions by R3 and R4. (IC1 must be 4011A). Outputs from gates are combined by op amp IC2, which provides low impedance output.

FOUR CHANNEL MIXER



+9V to +24V
Output
.1μf
1, 14
9
LM381
7, 8
A
1μf
500KΩ
2, 13
4
22KΩ
B
1μf
500KΩ
820Ω
2.2KΩ
C
1μf
500KΩ
D
1μf
500KΩ

Fig. 6-10

Circuit Notes

High gain op amp combines up to four individually controlled input signals. The dc power source should be well filtered (battery is ideal), and the circuit should be well shielded to prevent hum pickup.

7

Audio Oscillators

The sources of the following circuits are contained in the Sources section beginning on page 730. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

Wien Bridge Oscillator
Wien Bridge Oscillator
Wien Bridge Oscillator
Very Low Frequency Generator
Audio Oscillator
Sine Wave Oscillator
Easily Tuned Sine/Square Wave Oscillators
Wien Bridge Sine Wave Oscillator
Phase Shift Oscillator
Tone Encoder
Feedback Oscillator
Phase Shift Oscillator
800 Hz Oscillator
Tunable Single Comparator Oscillator
Wide Range Oscillator (Frequency Range of 500 to 1)
Wien Bridge Oscillator
Wien Bridge Sine Wave Oscillator

WIEN BRIDGE OSCILLATOR

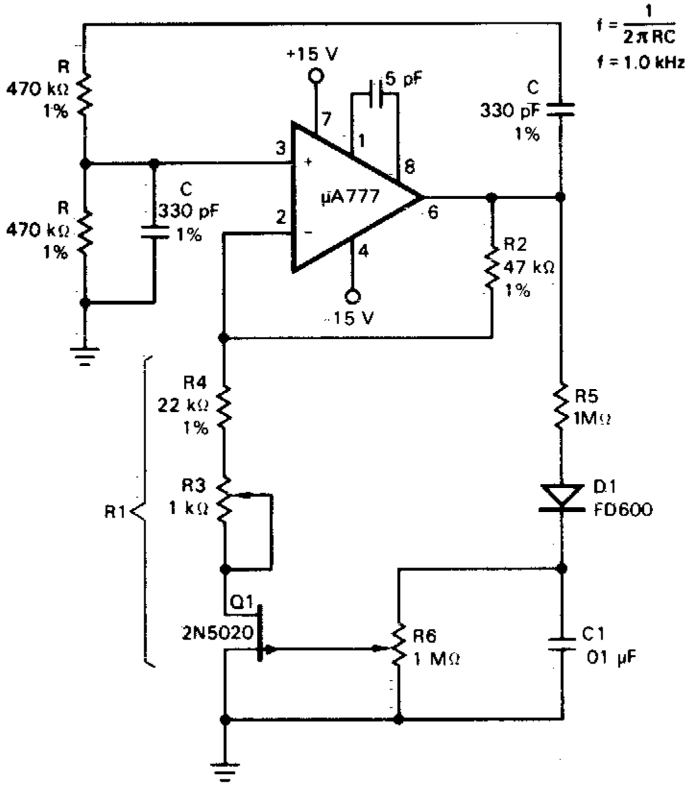

$f = \frac{1}{2\pi RC}$
f = 1.0 kHz
R
470 kΩ
1%
R
470 kΩ
1%
C
330 pF
1%
C
330 pF
1%
+15 V
5 pF
μA777
-15 V
R2
47 kΩ
1%
R4
22 kΩ
1%
R3
1 kΩ
R1
Q1
2N5020
R5
1MΩ
D1
FD600
R6
1 MΩ
C1
.01 μF
1
2
3
4
6
7
8

Fig. 7-1

Circuit Notes

Field effect transistor, Q1, operates in the linear resistive region to provide automatic gain control. Because the attenuation of the RC network is one-third at the zero phase-shift oscillation frequency, the amplifier gain determined by resistor R2 and equivalent resistor R1 must be just equal to three to make up the unity gain positive feedback requirement needed for stable oscillation. Resistors R3 and R4 are set to approximately 1000 ohm less than the required R1 resistance. The FET dynamically provides the trimming resistance needed to make R1 one-half of the resistance of R2. The circuit composed of R5, D1, and C1 isolates, rectifies, and filters the output sine wave, converting it into a dc potential to control the gate of the FET. For the low drain-to-source voltages used, the FET provides a symmetrical linear resistance for a given gate-to-source voltage.

WIEN BRIDGE OSCILLATOR

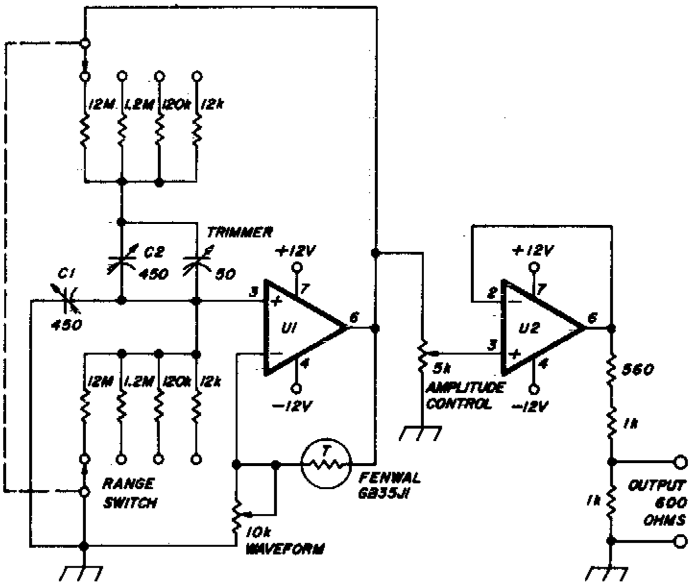
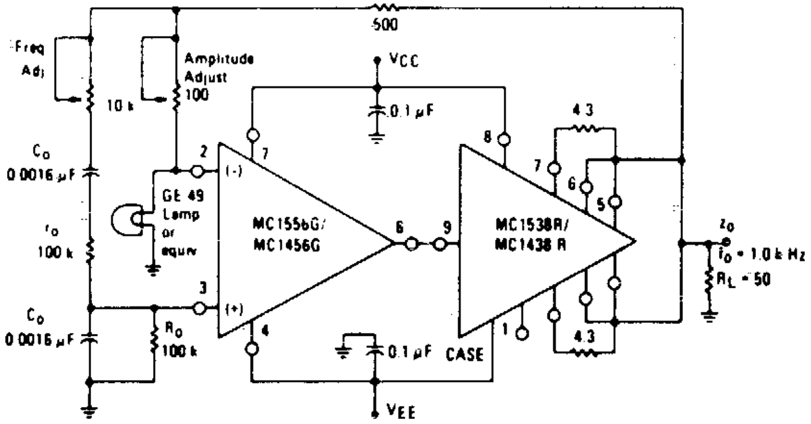

12M
1.2M
120k
12k
TRIMMER
C2
450
50
+12V
C1
450
3
7
U1
6
4
−12V
12M
1.2M
120k
12k
RANGE SWITCH
10k WAVEFORM
T
FENWAL GB35J1
5k AMPLITUDE CONTROL
+12V
2
7
U2
6
3
4
−12V
560
1k
1k
OUTPUT 600 OHMS

Fig. 7-2

Circuit Notes

Wien bridge sine-wave oscillator using two RCA CA3140 op amps covers 30 Hz to 100 kHz with less than 0.5 percent total harmonic distortion. The 10k pot is adjusted for the best waveform. Capacitor C1 and C2 are a two-gang, 450-pF variable with its frame isolated from ground. Maximum output into a 600-ohm load is about 1 volt rms.

WIEN BRIDGE OSCILLATOR

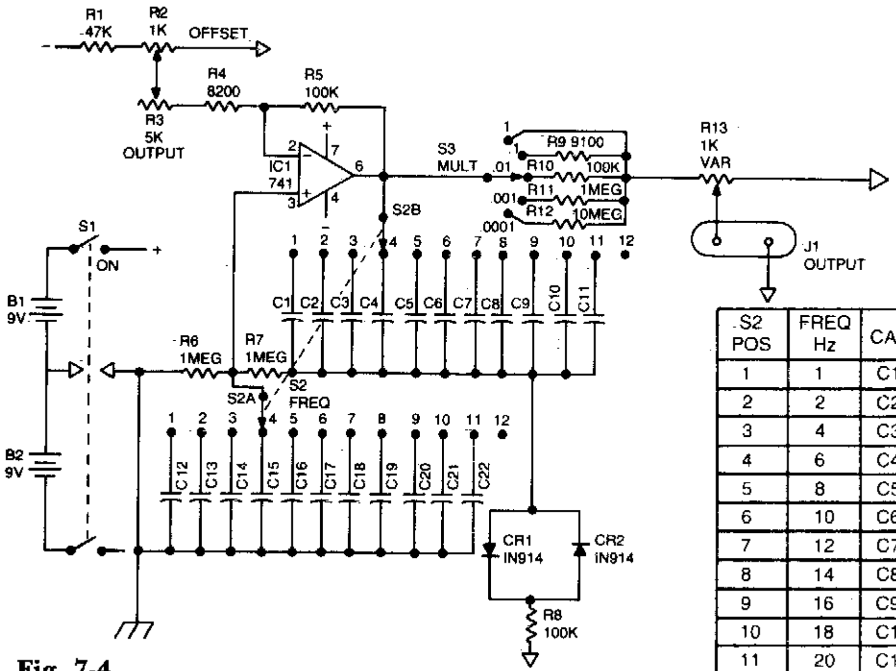

500
Freq Adj
10 k
Amplitude Adjust
100
VCC
0.1 μF
4.3
C0
0.0016 μF
2
(−)
7
8
7
6
5
GE 49 Lamp or equiv
MC1556G/
MC1456G
6
9
MC1538R/
MC1438R
r0
100 k
3
(+)
z0
f0 = 1.0 kHz
RL = 50
C0
0.0016 μF
R0
100 k
4
0.1 μF
CASE
1
4.3
VEE

Characteristics

I. $f_0 = \frac{1}{2\pi r_0 C_0}$ from 0.01 Hz to 10 kHz II. IOS = 200 mA

Fig. 7-3

VERY LOW FREQUENCY GENERATOR


R1 47K
R2 1K
OFFSET
R4 8200
R5 100K
R3 5K OUTPUT
IC1 741
S3 MULT
R9 9100
R10 100K
R11 1MEG
R12 10MEG
.01
.001
.0001
R13 1K VAR
J1 OUTPUT
S2B
S1
ON
B1 9V
B2 9V
R6 1MEG
R7 1MEG
S2A
S2 FREQ
C1 C2 C3 C4 C5 C6 C7 C8 C9 C10 C11
C12 C13 C14 C15 C16 C17 C18 C19 C20 C21 C22
CR1 IN914
CR2 IN914
R8 100K

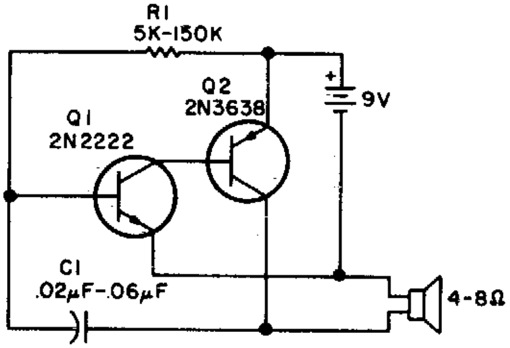
S2 POS	FREQ Hz	CAPACITOR	VALUE μF
1	1	C1=C12=	.5 + 1
2	2	C2=C13=	.15 + J5
3	4	C3=C14=	.15
4	6	C4=C15=	.1
5	8	C5=C16=	.066 + .0068
6	10	C6=C17=	.D5 + D1
7	12	C7=C 18=	.05
8	14	C8=C19=	.033 +.01
9	16	C9=C20=	0.33 +.0047
10	18	C10=C21=	.033
11	20	C11=C22=	.015+.015
12	–	–	–

Fig. 7-4

Circuit Notes

Wien bridge oscillator generates frequencies of 1 Hz and 2 to 20 Hz in 2 Hz steps. Maximum output amplitude is 3 volts rms of 8.5 volts peak-to-peak. A pot-and-switch attenuator allows the output level to be set with a fair degree of precision to any value within a range of 5 decades.

AUDIO OSCILLATOR


R1 5K-150K
Q2 2N3638
Q1 2N2222
+ 9V
C1 .02μF-.06μF
4-8Ω

Circuit Notes

Almost any transistor will work.
R1 and C1 will vary the tone.

Fig. 7-5

SINE WAVE OSCILLATOR

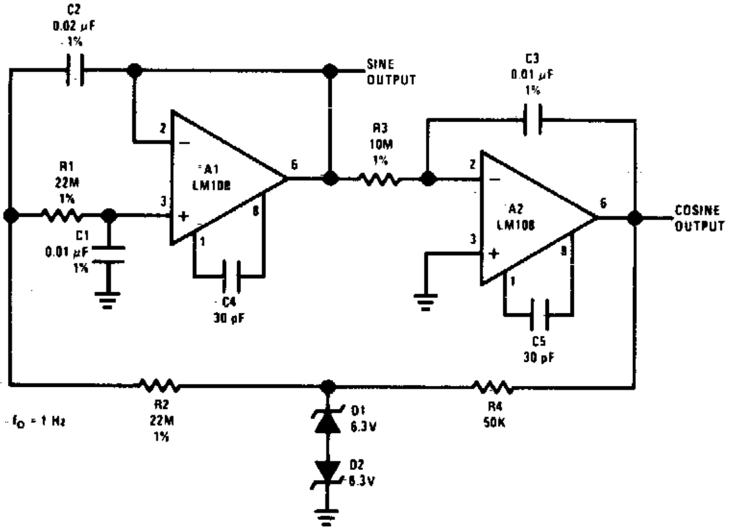
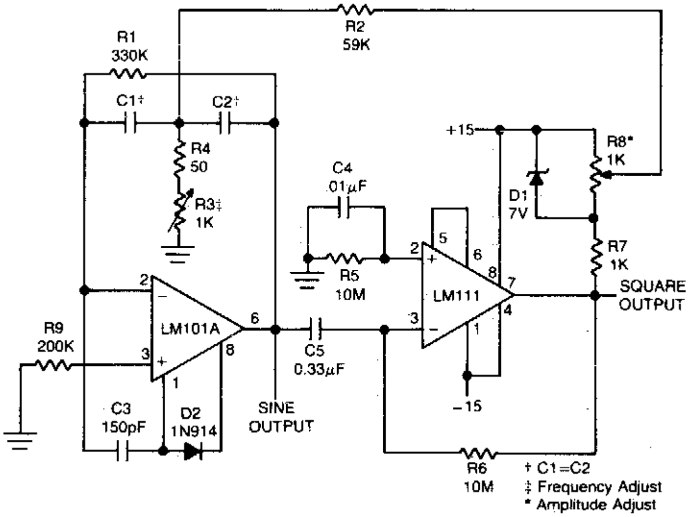

C2
0.02 µF
1%
SINE OUTPUT
C3
0.01 µF
1%
R3
10M
1%
R1
22M
1%
A1
LM108
A2
LM108
COSINE OUTPUT
C1
0.01 µF
1%
C4
30 pF
C5
30 pF
$f_0 = 1$ Hz
R2
22M
1%
D1
6.3V
D2
6.3V
R4
50K

Fig. 7-6

Circuit Notes

The oscillator delivers a high-purity sinusoid with a stable frequency and amplitude.

EASILY TUNED SINE/SQUARE WAVE OSCILLATORS


R1
330K
R2
59K
C1†
C2†
R4
50
R3‡
1K
+15
R8*
1K
D1
7V
C4
.01µF
R5
10M
R7
1K
LM111
SQUARE OUTPUT
R9
200K
LM101A
C5
0.33µF
C3
150pF
D2
1N914
SINE OUTPUT
−15
R6
10M
† C1=C2
‡ Frequency Adjust
* Amplitude Adjust

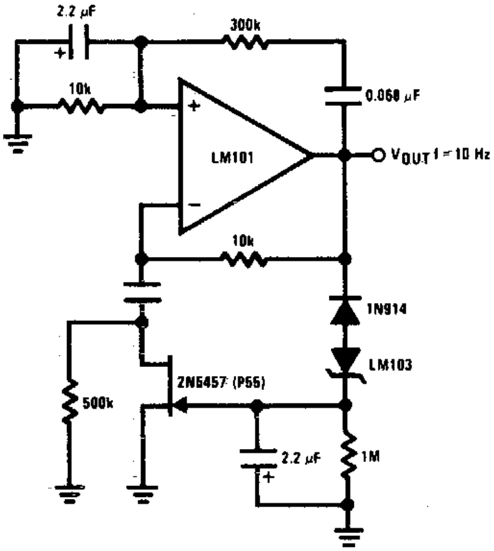
$$F_0 = \frac{1}{2\pi C_1 \sqrt{R_3 R_1}}$$

Fig. 7-7

Circuit Notes

This circuit will provide both a sine and square wave output for frequencies from below 20 Hz to above 20 kHz. The frequency of oscillation is easily tuned by varying a single resistor.

WIEN BRIDGE SINE WAVE OSCILLATOR


2.2 μF
+
300k
10k
0.068 μF
+
LM101
−
V_{OUT} f = 10 Hz
10k
1N914
LM103
2N5457 (P55)
500k
2.2 μF
+
1M

Peak output voltage
$V_p \cong V_z + 1V$

Fig. 7-8

Circuit Notes

Using the 2N5457 JFET as a voltage variable resistor in the amplifier feedback loop, produces a low distortion, constant amplitude sine wave getting the amplifier loop gain just right. The LM103 zener diode provides the voltage reference for the peak sine wave amplitude.

PHASE-SHIFT OSCILLATOR

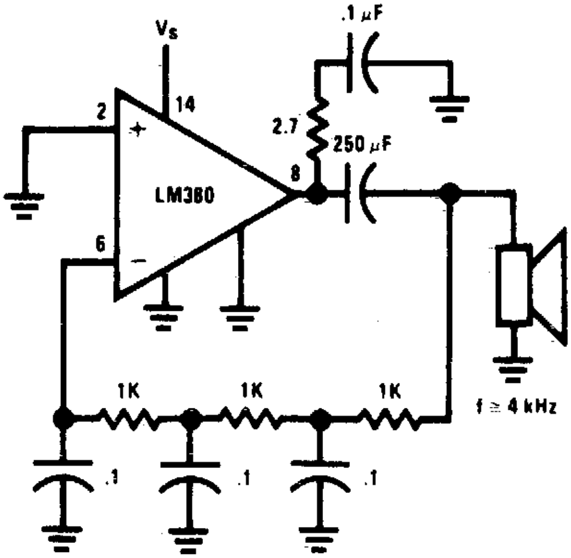

V_s
.1 μF
14
2
+
2.7
250 μF
8
LM380
6
−
1K
1K
1K
f ≅ 4 kHz
.1
.1
.1

Fig. 7-9

Circuit Notes

Circuit uses a simple RC network to produce an exceptionally shrill tone from a miniature speaker. With the parts values shown, the circuit oscillates at a frequency of 3.6 kHz and drives a miniature 2½" speaker with ear-piercing volume. The output waveform is a square wave with a width of 150 μs, sloping rise and fall times, and a peak-to-peak amplitude of 4.2 volts (when powered by 9 volts). Current drain of the oscillator is 90 mA at 9 volts, and total power dissipation at this voltage is 0.81 watt, which is well below the 1.25 watts the 14-pin version will absorb (at room temperature) before shutting down.

TONE ENCODER

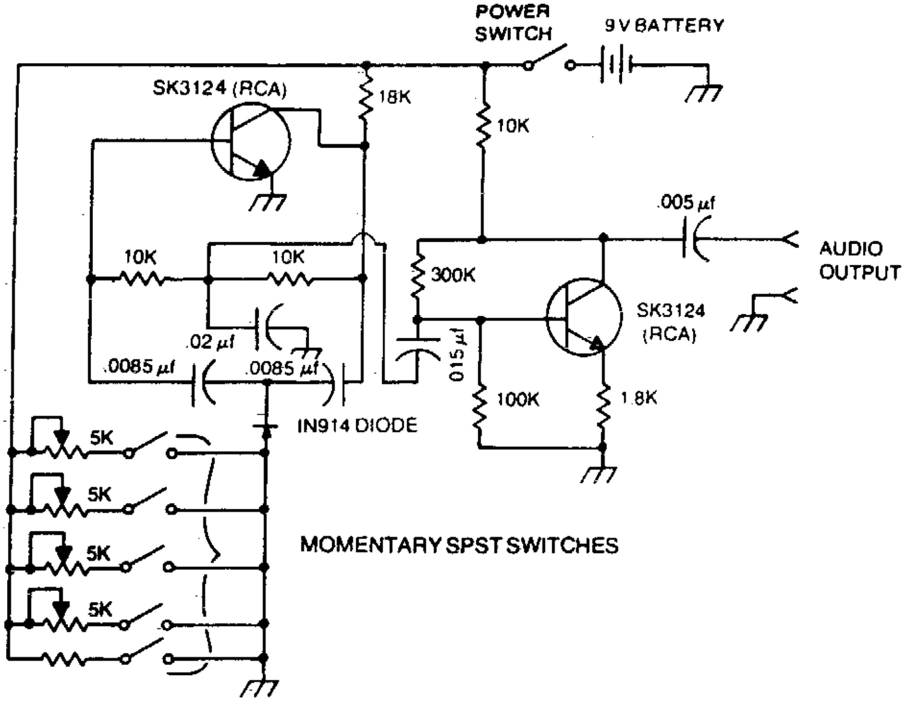
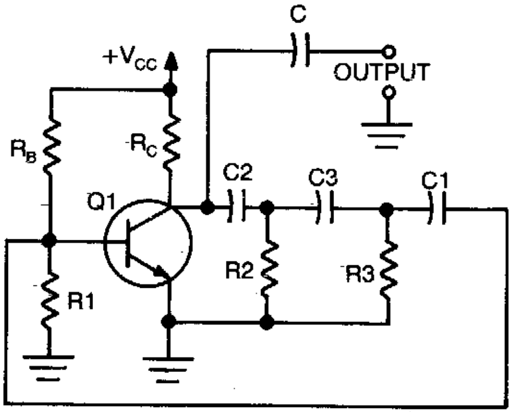

POWER SWITCH
9V BATTERY
SK3124 (RCA)
18K
10K
.005 µf
AUDIO OUTPUT
10K
10K
300K
SK3124 (RCA)
.02 µf
.0085 µf
.0085 µf
.015 µf
100K
1.8K
IN914 DIODE
5K
5K
5K
5K
MOMENTARY SPST SWITCHES

Fig. 7-10

Circuit Notes

A basic twin-T circuit uses resistors for accurately setting the frequency of the output tones, selected by pushbutton. Momentary switches produce a tone only when the button is depressed.

FEEDBACK OSCILLATOR


C
$+V_{CC}$
OUTPUT
R_B
R_C
C2
C3
C1
Q1
R2
R3
R1

Circuit Notes

Circuit oscillates because the transistor shifts the phase of the signal 180° from the base to the collector. Each of the RC networks in the circuit is designed to shift the phase 60° at the frequency of oscillation for a total of 180°. The appropriate values of R and C for each network is found from $f = 1/2\sqrt{3}\pi RC)$; that equation allows for the 60° phase shift required by the design.

Fig. 7-11

PHASE SHIFT OSCILLATOR

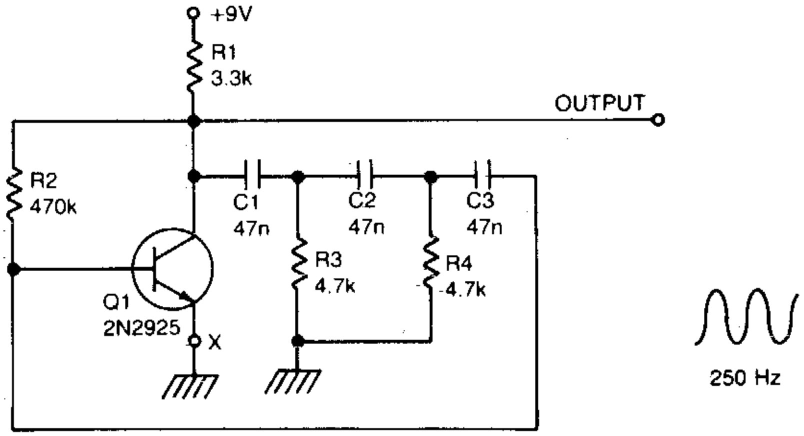

+9V
R1
3.3k
OUTPUT
R2
470k
C1
47n
C2
47n
C3
47n
R3
4.7k
R4
4.7k
Q1
2N2925
X
250 Hz

Fig. 7-12

Circuit Notes

A single transistor makes a simple phase shift oscillator. The output is a sine wave with distortion of about 104. The sine wave purity can be increased by putting a variable resistor (25 ohms) in the emitter lead of Q1 (x). The resistor is adjusted so the circuit is only just oscillating, then the sine wave is relatively pure. Operating frequency may be varied by putting a 10 K variable resistor in series with R3, or by changing C1, C2, and C3. Making C1, 2, 3 equal to 100 nF will halve the operating frequency. Operating frequency can also be voltage controlled by a FET in series with R3, or optically controlled by an LDR in series with R3.

800 Hz OSCILLATOR

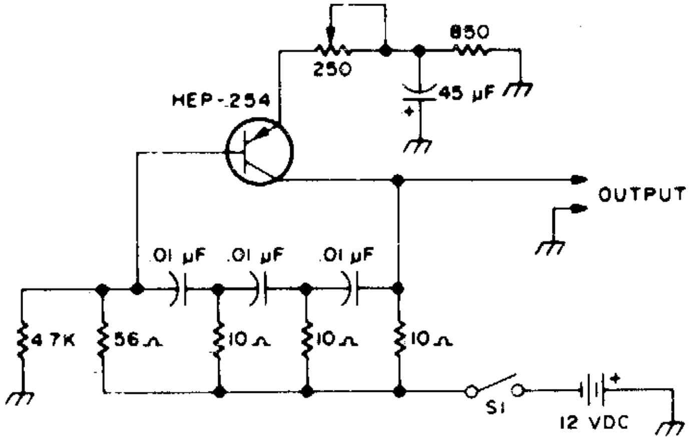
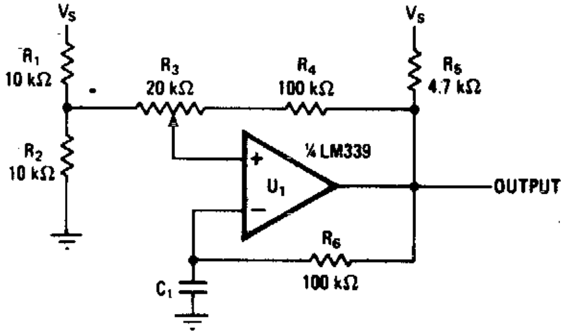

850
250
HEP-254
45 µF
+
OUTPUT
.01 µF
.01 µF
.01 µF
4.7K
56Ω
10Ω
10Ω
10Ω
S1
+
12 VDC

Fig. 7-13

Circuit Notes

The following transistors may be used: HEP-254, O.C-2, SK-3004, AT30H. To increase the frequency, decrease the value of the capacitors in the ladder network.

TUNABLE SINGLE COMPARATOR OSCILLATOR



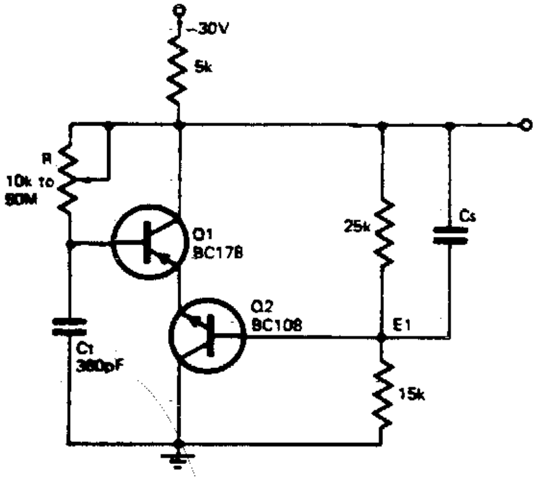
V_S
R_1
10 kΩ
R_2
10 kΩ
R_3
20 kΩ
R_4
100 kΩ
V_S
R_5
4.7 kΩ
¼ LM339
U_1
+
−
OUTPUT
R_6
100 kΩ
C_1

Fig. 7-14

Circuit Notes

Varying the amount of this comparator circuit's hysteresis makes it possible to vary output frequencies in the 740-Hz to 2.7-kHz range smoothly. The amount of hysteresis together with time constant R6C1 determines how much time it takes for C1 to charge or discharge to the new threshold after the output voltage switches.

WIDE RANGE OSCILLATOR (FREQUENCY RANGE OF 5000 TO 1)



−30V
5k
R
10k to 50M
Q1
BC178
Q2
BC108
25k
Cs
E1
Ct
380pF
15k

Fig. 7-15

Circuit Notes

Timing resistor R may be adjusted to any value between 10 K and 50 M to obtain a frequency range from 400 kHz to 100 Hz. Returning the timing resistor to the collector of Q1 ensures that Q1 draws its base current only from the timing capacitor Ct. The timing capacitor recharges when the transistors are off, to a voltage equal to the base emitter voltage of Q2 plus the base emitter drops of Q1 and Q2. The transistors then start into conduction. Capacitor Cs is used to speed up the transition. A suitable value would be in the region of 100 pF.

WIEN BRIDGE OSCILLATOR

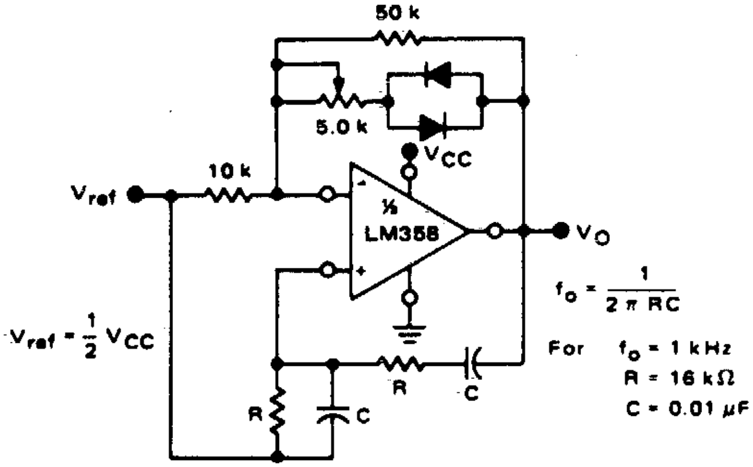

50 k
5.0 k
Vcc
10 k
Vref
½
LM358
Vo
fo = 1 / 2π RC
Vref = 1/2 Vcc
For fo = 1 kHz
R = 16 kΩ
C = 0.01 μF
R
C
R
C

Fig. 7-16

WIEN BRIDGE SINE WAVE OSCILLATOR

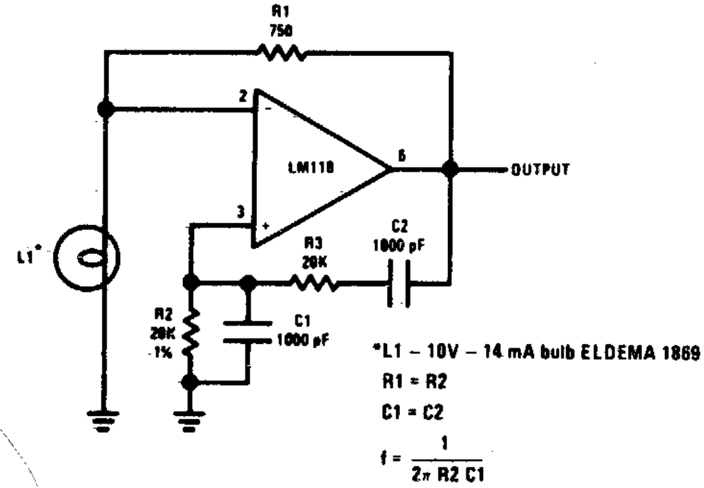

R1
750
2
LM118
6
OUTPUT
3
L1*
R3
20K
C2
1000 pF
R2
20K
1%
C1
1000 pF
*L1 – 10V – 14 mA bulb ELDEMA 1869
R1 = R2
C1 = C2
f = 1 / 2π R2 C1

Fig. 7-17

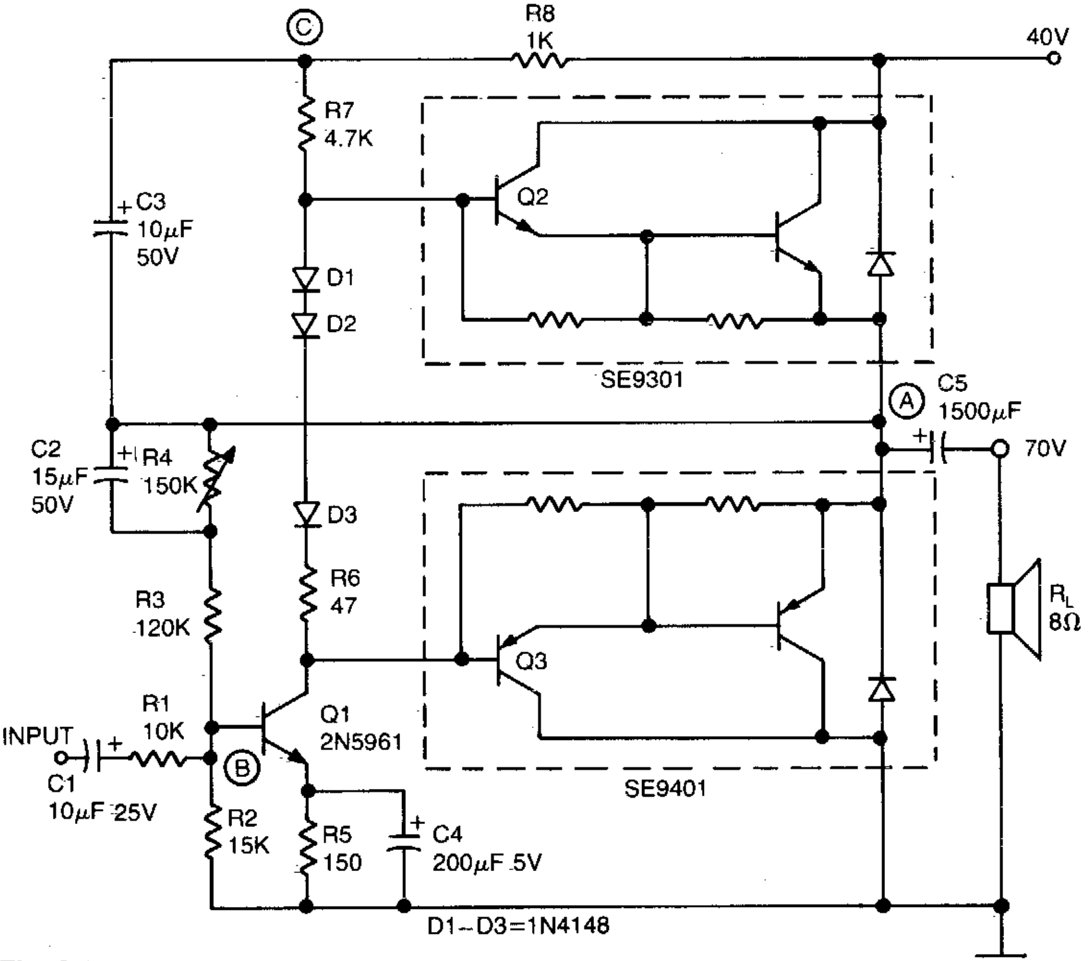
8

Audio Power Amplifiers

The sources of the following circuits are contained in the Sources section beginning on page 730. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

Low Cost 20 W Audio Amplifier
75 Watt Audio Amplifier with Load Line Protection
Bridge Amplifier
Noninverting Amplifier Using Single Supply
Noninverting Amplifier Using Split Supply
6 W, 8 Ohm Output Transformerless Amplifier
12 W Low-Distortion Power Amplifier
10 W Power Amplifier
Stereo Amplifier with Av = 200
AM Radio Power Amplifier
470 mW Complementary-Symmetry Audio Amplifier
Novel Loudspeaker Coupling Circuit
Noninverting Ac Power Amplifier
Inverting Power Amplifier
Noninverting Power Amplifier
4 W Bridge Amplifier
Phono Amplifier with a "Common Mode" Volume and Tone with Control
Phono Amplifier
Phonograph Amplifier (Ceramic Cartridge)
Inverting Unity Gain Amplifier
Bridge Audio Power Amplifier
Phono Amplifier
High Slew Rate Power Op Amp/Audio Amp
16 W Bridge Amplifier

LOW COST 20 W AUDIO AMPLIFIER



C
R8
1K
40V
R7
4.7K
Q2
+C3
10μF
50V
D1
D2
SE9301
A
C5
1500μF
70V
C2
15μF
50V
+
R4
150K
D3
R6
47
R3
120K
RL
8Ω
Q3
R1
10K
Q1
2N5961
INPUT
+
C1
10μF 25V
B
SE9401
R2
15K
R5
150
+
C4
200μF 5V
D1–D3=1N4148

Fig. 8-1

Circuit Notes

This simple inexpensive audio amplifier can be constructed using a couple of TO-220 monolithic Darlington transistors for the push-pull output stage. Frequency response is flat within 1 dB from 30 Hz to 200 kHz with typical harmonic distortion below 0.2%. The amplifier requires only 1.2 V_{rms} for a full 20 W output into an 8 ohm load. Only one other transistor is needed, the TO-92 low-noise high-gain 2N5961 (Q1), to provide voltage gain for driving the output Darlingtons. Its base (point B) is the tie point for ac and dc feedback as well as for the signal input. Input resistance is 10 K. The center voltage at point A is set by adjusting resistor R4. A bootstrap circuit boosts the collector supply voltage of Q1 (point C) to ensure sufficient drive voltage for Q2. This also provides constant voltage across R7, which therefore acts as a current source and, together with diodes D1-D3, reduces low-signal crossover distortion.

75 WATT AUDIO AMPLIFIER WITH LOAD LINE PROTECTION

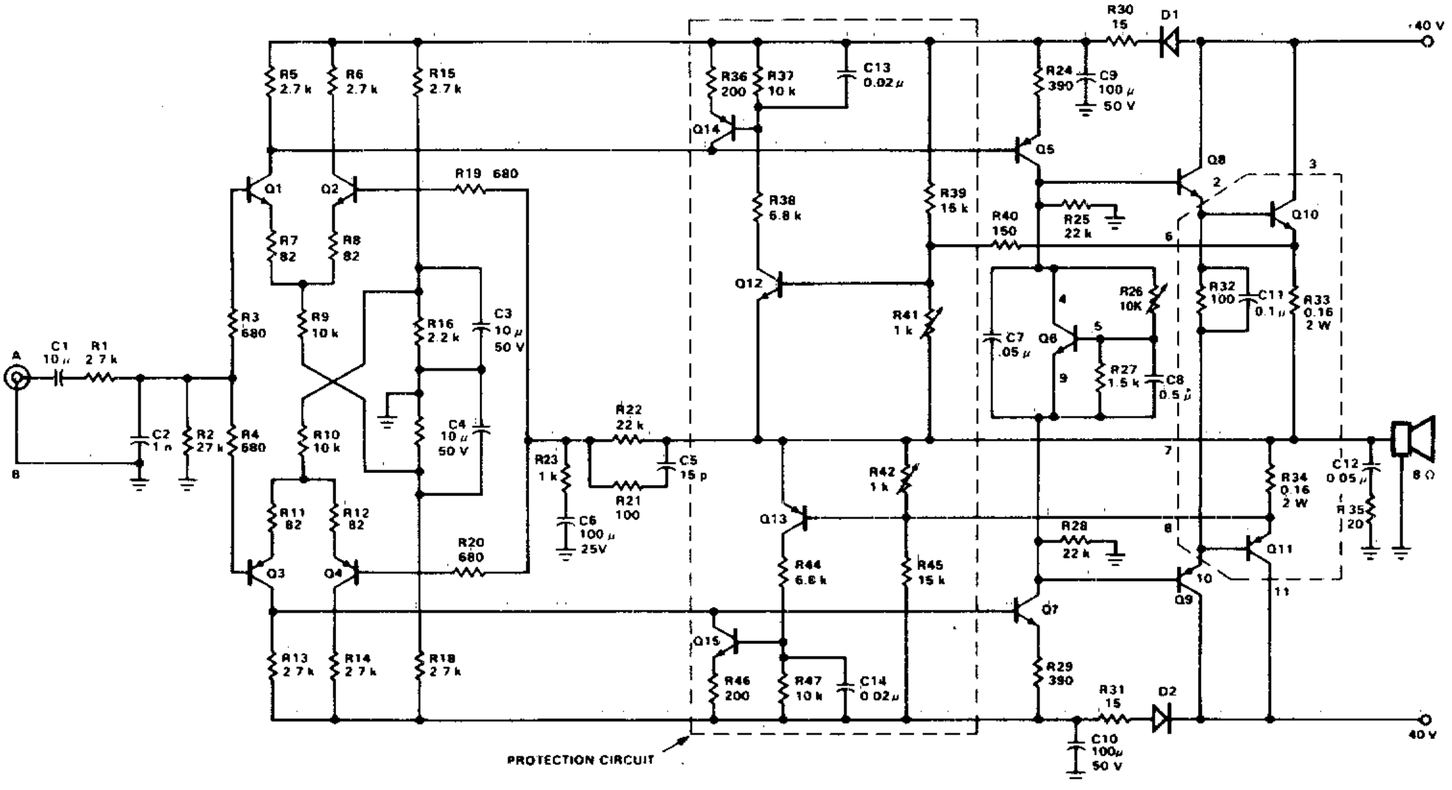
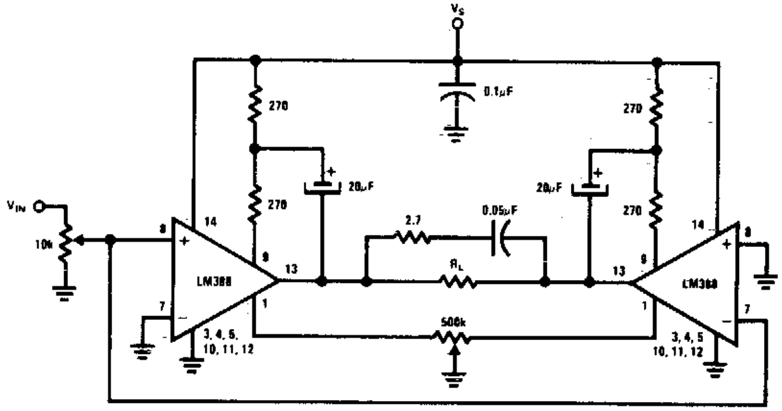

A
B
C1 10 μ
R1 2.7 k
C2 1 n
R2 27 k
R3 680
R4 680
R5 2.7 k
R6 2.7 k
R7 82
R8 82
R9 10 k
R10 10 k
R11 82
R12 82
R13 2.7 k
R14 2.7 k
R15 2.7 k
R16 2.2 k
R18 2.7 k
R19 680
R20 680
C3 10 μ 50 V
C4 10 μ 50 V
Q1
Q2
Q3
Q4
R22 22 k
R21 100
R23 1 k
C5 15 p
C6 100 μ 25V
R36 200
R37 10 k
C13 0.02 μ
Q14
R38 6.8 k
Q12
R39 15 k
R41 1 k
R42 1 k
Q13
R44 6.8 k
R45 15 k
Q15
R46 200
R47 10 k
C14 0.02 μ
PROTECTION CIRCUIT
R24 390
C9 100 μ 50 V
R30 15
D1
-40 V
Q5
R40 150
R25 22 k
Q6
C7 .05 μ
R26 10K
R27 1.5 k
C8 0.5 μ
Q7
R28 22 k
R29 390
R31 15
D2
C10 100μ 50 V
40 V
Q8
Q9
Q10
Q11
R32 100
C11 0.1 μ
R33 0.16 2 W
R34 0.16 2 W
C12 0.05 μ
R35 20
8 Ω
2
3
4
5
6
7
8
9
10
11

Fig. 8-2

BRIDGE AMPLIFIER


V_S
270
0.1µF
270
20µF
20µF
270
270
V_{IN}
10k
8
14
2.7
0.05µF
14
8
+
9
R_L
9
LM388
13
13
LM388
7
1
500k
1
7
−
3, 4, 5,
10, 11, 12
3, 4, 5
10, 11, 12

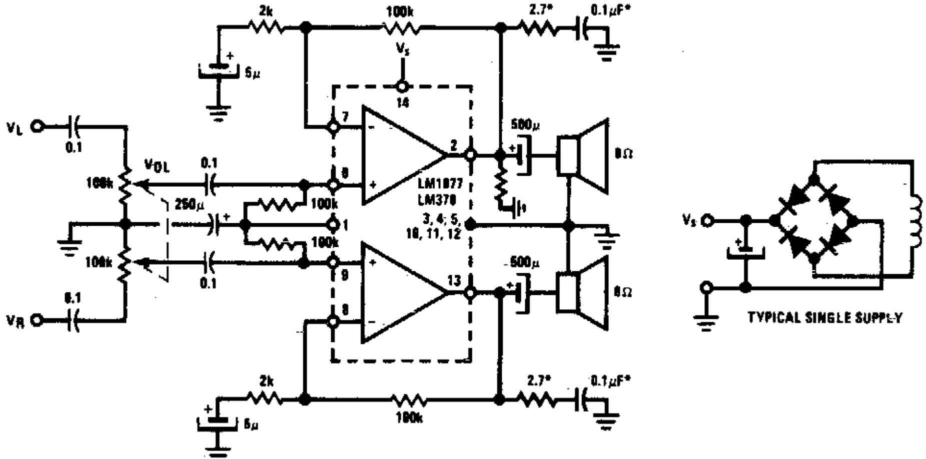
V_S = 6V R_L = 4Ω P_O = 1.0W
V_S = 12V R_L = 8Ω P_O = 3.5W

Fig. 8-3

Circuit Notes

This circuit is for low voltage applications requiring high power outputs. Output power levels of 1.0 W into 4 ohm from 6 V and 3.5 V into 8 ohm from 12 V are typical. Coupling capacitors are not necessary since the output dc levels will be within a few tenths of a volt of each other. Where critical matching is required the 500 K potentiometer is added and adjusted for zero dc current flow through the load.

NONINVERTING AMPLIFIER USING SINGLE SUPPLY


2k
100k
2.7*
0.1µF*
V_S
5µ
14
V_L
0.1
7
500µ
V_{OL}
0.1
2
8Ω
100k
8
LM1877
LM378
250µ
100k
1
3, 4, 5,
10, 11, 12
100k
100k
9
500µ
0.1
13
8Ω
0.1
8
V_R
V_S
TYPICAL SINGLE SUPPLY
2k
2.7*
0.1µF*
100k
5µ

* (LM1877 ONLY)

6 – RESISTORS
9 – CAPACITORS

Fig. 8-4

NONINVERTING AMPLIFIER USING SPLIT SUPPLY

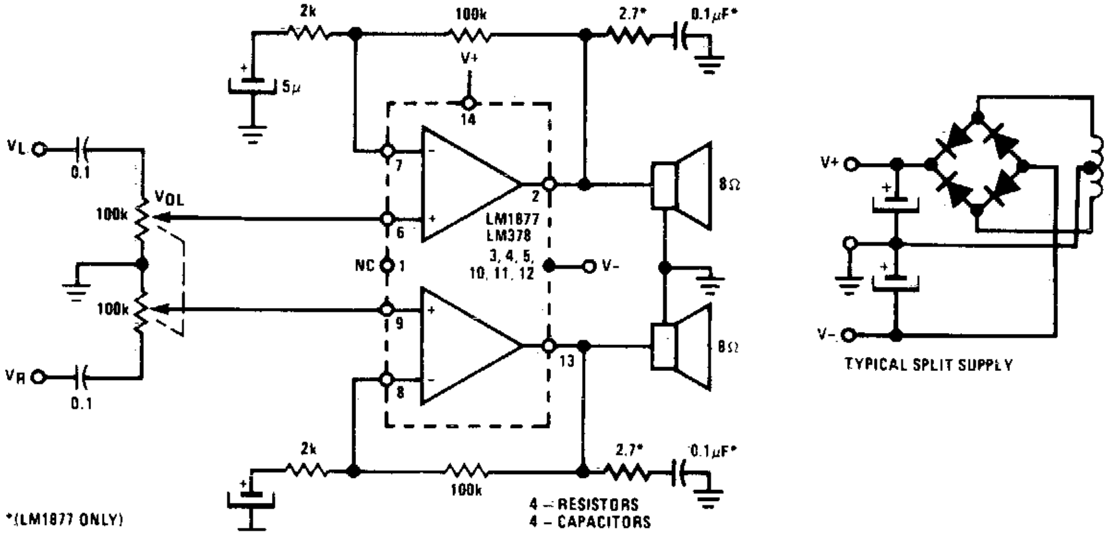
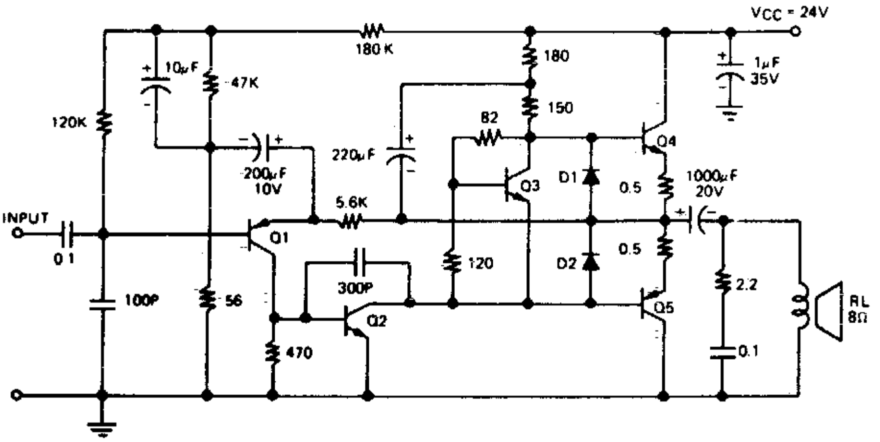

2k
100k
2.7*
0.1μF*
V+
5μ
14
VL
0.1
VOL
100k
7
6
2
8Ω
LM1877
LM378
NC 1
3, 4, 5,
10, 11, 12
V-
100k
9
8
13
8Ω
VR
0.1
2k
100k
2.7*
0.1μF*
4 – RESISTORS
4 – CAPACITORS
*(LM1877 ONLY)
V+
V-
TYPICAL SPLIT SUPPLY

Fig. 8-5

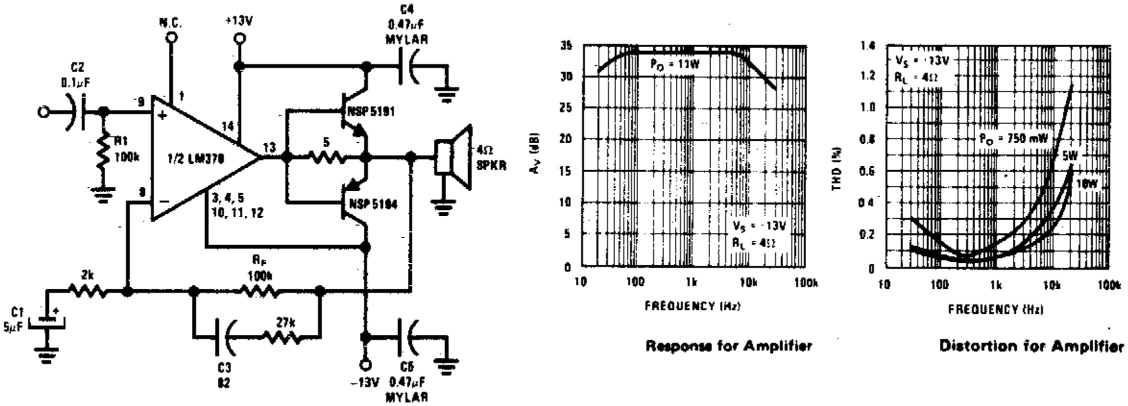
6 W,8 'OHM OUTPUT TRANSFORMERLESS AMPLIFIER


VCC = 24V
180 K
180
1μF
35V
10μF
47K
120K
150
82
Q4
200μF
10V
220μF
Q3
D1
0.5
1000μF
20V
INPUT
0.1
5.6K
Q1
120
D2
0.5
100P
56
300P
2.2
RL
8Ω
Q2
Q5
470
0.1

Q1	NB021EY
Q2	NB211YY
Q3	NR001E
Q4	NA41U
Q5	NA42U

Fig. 8-6

12 W LOW-DISTORTION POWER AMPLIFIER

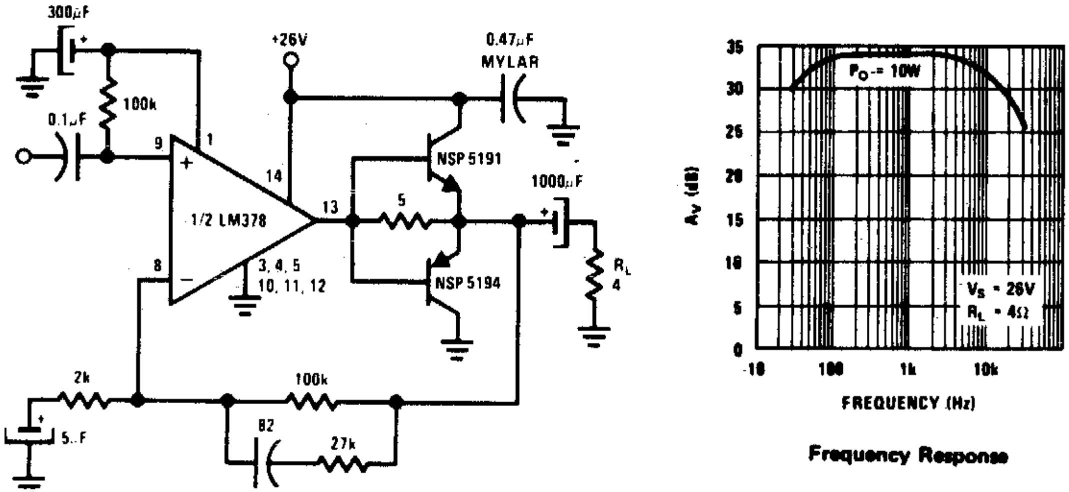

N.C.
+13V
C4
0.47μF
MYLAR
C2
0.1μF
9
1
14
+
R1
100k
1/2 LM378
13
5
NSP 5191
4Ω
SPKR
8
−
3, 4, 5
10, 11, 12
NSP 5194
R_F
100k
2k
C1
5μF
27k
C3
82
−13V
C5
0.47μF
MYLAR
A_V (dB)
P_O = 11W
V_S = ±13V
R_L = 4Ω
FREQUENCY (Hz)
THD (%)
V_S = ±13V
R_L = 4Ω
P_O = 750 mW
5W
10W
FREQUENCY (Hz)

Response for Amplifier

Distortion for Amplifier

Fig. 8-7

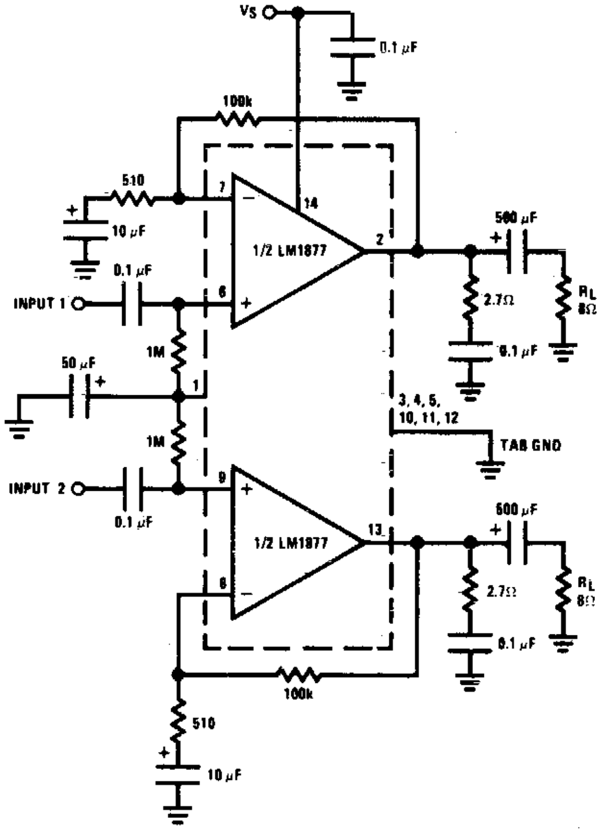
10 W POWER AMPLIFIER


300μF
100k
0.1μF
9
1
+26V
0.47μF
MYLAR
+
14
1/2 LM378
13
5
NSP 5191
1000μF
8
−
3, 4, 5
10, 11, 12
NSP 5194
R_L
4
2k
5μF
100k
82
27k
A_V (dB)
P_O = 10W
V_S = 26V
R_L = 4Ω
FREQUENCY (Hz)

Frequency Response

Fig. 8-8

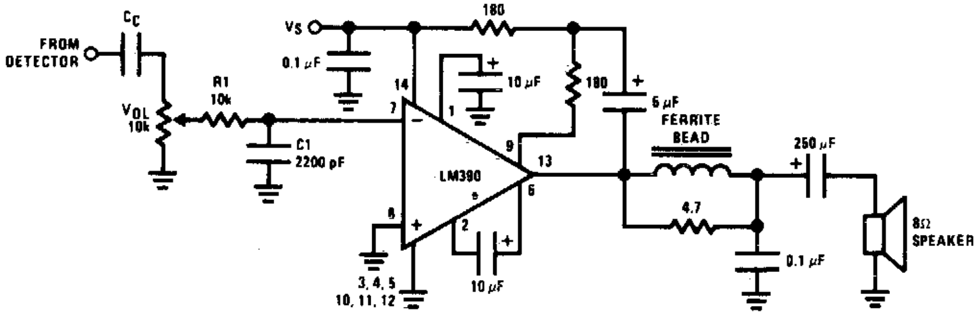
STEREO AMPLIFIER WITH Av = 200



Vs
0.1 µF
100k
510
10 µF
0.1 µF
INPUT 1
7
14
1/2 LM1877
2
6
500 µF
2.7Ω
RL
8Ω
0.1 µF
1M
50 µF
1
3, 4, 5,
10, 11, 12
TAB GND
1M
INPUT 2
9
0.1 µF
1/2 LM1877
13
500 µF
8
2.7Ω
RL
8Ω
0.1 µF
100k
510
10 µF

Fig. 8-9

AM RADIO POWER AMPLIFIER



CC
FROM
DETECTOR
VOL
10k
R1
10k
C1
2200 pF
Vs
0.1 µF
14
7
1
10 µF
180
180
5 µF
FERRITE
BEAD
9
13
LM390
6
250 µF
4.7
8
2
10 µF
0.1 µF
3, 4, 5
10, 11, 12
8Ω
SPEAKER

Note 1: Twist supply lead and supply ground very tightly.

Note 2: Twist speaker lead and ground very tightly.

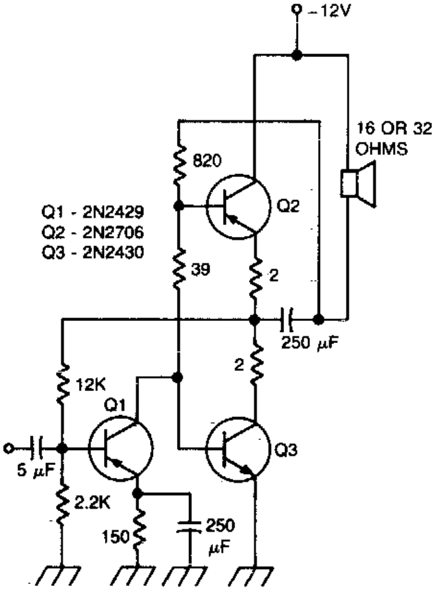
Note 3: Ferrite bead is Ferroxcube K5-001-001/3B with 3 turns of wire.

Note 4: R1C1 band limits input signals.

Note 5: All components must be spaced very close to IC.

Fig. 8-10

470 mW COMPLEMENTARY-SYMMETRY AUDIO AMPLIFIER



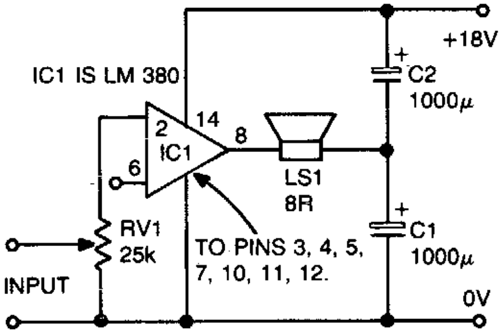
-12V
16 OR 32
OHMS
820
Q2
Q1 - 2N2429
Q2 - 2N2706
Q3 - 2N2430
39
2
250 μF
2
12K
Q1
Q3
5 μF
2.2K
150
250
μF

Fig. 8-11

Circuit Notes

This circuit has less than 2% distortion and is flat within 3 dB from 15 Hz to 130 kHz.

NOVEL LOUDSPEAKER COUPLING CIRCUIT



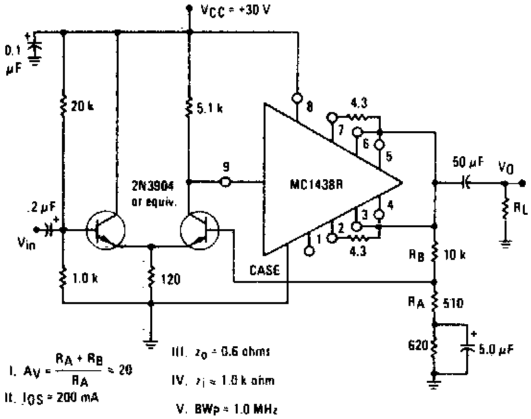
+18V
IC1 IS LM 380
C2
1000μ
2
14
8
IC1
6
LS1
8R
C1
1000μ
RV1
25k
TO PINS 3, 4, 5,
7, 10, 11, 12.
INPUT
0V

Fig. 8-12

Circuit Notes

The ground side of the speaker is connected to the junction of two equal high value capacitors (1000 μF is typical) across the supply. The amplifier output voltage will be $V_s/2$, and so will the voltage across C1 (if C1 and C2 are equal); so as the supply voltage builds up, the dc voltage across the speaker will remain zero, eliminating the switch-on surge. C1 and C2 will also provide supply smoothing. The circuit is shown with the LM380, but could be applied to any amplifier circuit, providing that the dc voltage at the output is half the supply voltage.

NONINVERTING AC POWER AMPLIFIER


V_{CC} = +30 V
0.1 μF
20 k
5.1 k
2N3904 or equiv.
9
8
7
6
5
4
3
2
1
4.3
4.3
MC1438R
50 μF
V_O
R_L
.2 μF
V_{in}
1.0 k
120
CASE
R_B
10 k
R_A
510
620
5.0 μF

I. $A_V = \frac{R_A + R_B}{R_A} \approx 20$

II. I_{OS} = 200 mA

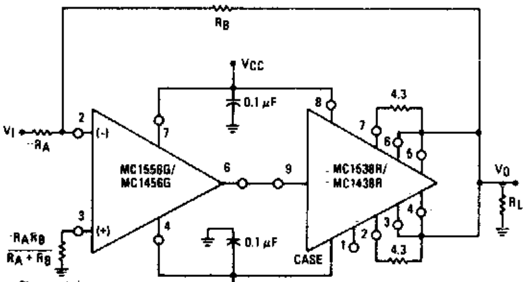
III. z_o = 0.6 ohms

IV. $z_i \approx$ 1.0 k ohm

V. $BW_P \approx$ 1.0 MHz

Fig. 8-13

INVERTING POWER AMPLIFIER


R_B
V_{CC}
0.1 μF
V_I
R_A
2
(-)
7
8
MC1556G/ MC1456G
6
9
MC1538R/ MC1438R
4.3
7
6
5
4
3
2
1
4.3
V_O
R_L
3
(+)
4
0.1 μF
CASE
$\frac{R_A R_B}{R_A + R_B}$

Characteristics

I. $A_V = \frac{-R_B}{R_A}$

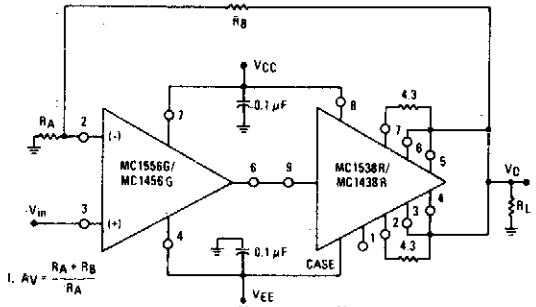
II. $z_o \approx \frac{R_A + R_B}{-R_A} 10^{-4}\,\Omega$

III. $z_i \approx R_A$

IV. $I_{OS} \approx$ 200 mA

Fig. 8-14

NONINVERTING POWER AMPLIFIER


R_B
V_{CC}
0.1 μF
R_A
2
(-)
7
8
MC1556G/ MC1456G
6
9
MC1538R/ MC1438R
4.3
7
6
5
4
3
2
1
4.3
V_O
R_L
V_{in}
3
(+)
4
0.1 μF
CASE
V_{EE}

I. $A_V = \frac{R_A + R_B}{R_A}$

II. $z_o \approx \frac{R_A + R_B}{R_A} 10^{-4}\,\Omega$

III. z_i > 30 megohm for f < 20 Hz

IV. I_{OS} = 200 mA

Fig. 8-15

4 W BRIDGE AMPLIFIER

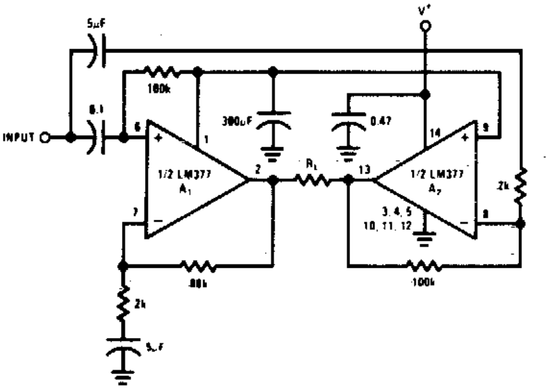
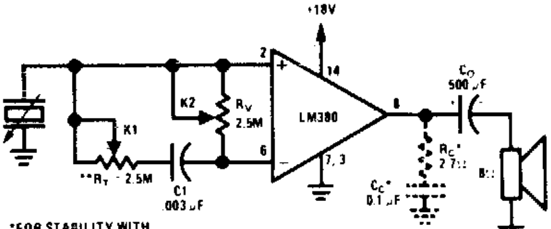

V^+
5μF
100k
0.1
INPUT
6
1
300μF
0.47
14
9
1/2 LM377 A_1
2
R_L
13
1/2 LM377 A_2
2k
7
3, 4, 5 10, 11, 12
8
100k
100k
2k
5μF

Fig. 8-16

PHONO AMPLIFIER WITH "COMMON MODE" VOLUME AND TONE CONTROL


+18V
2
14
6
7, 3
8
LM380
K1
K2
R_V
2.5M
**R_T - 2.5M
C1
.003 μF
C_O
500 μF
R_C*
2.7Ω
C_C*
0.1 μF
8Ω

*FOR STABILITY WITH HIGH CURRENT LOADS

**AUDIO TAPE POTENTIOMETER (10% OF R_T AT 50% ROTATION)

Fig. 8-17

PHONO AMPLIFIER

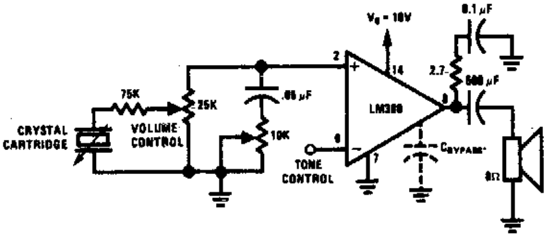

V_S = 18V
0.1 μF
2
14
6
7
8
LM380
2.7
500 μF
75K
25K
.05 μF
10K
CRYSTAL CARTRIDGE
VOLUME CONTROL
TONE CONTROL
C_{BYPASS}*
8Ω

Fig. 8-18

PHONOGRAPH AMPLIFIER (CERAMIC CARTRIDGE)

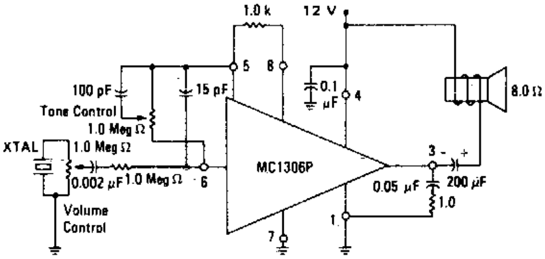

1.0 k
12 V
5
8
4
100 pF
15 pF
0.1 μF
Tone Control
1.0 Meg Ω
XTAL
1.0 Meg Ω
0.002 μF
1.0 Meg Ω
6
Volume Control
MC1306P
7
1
3
0.05 μF
200 μF
1.0
8.0 Ω

Fig. 8-19

INVERTING UNITY GAIN AMPLIFIER

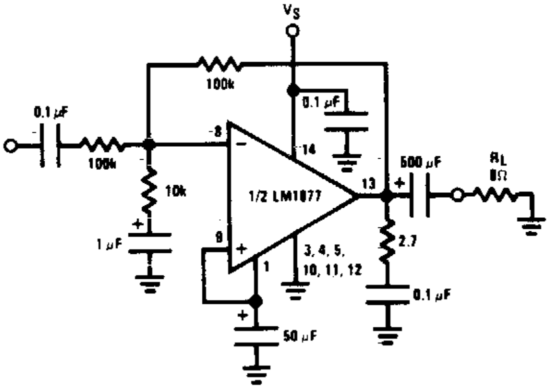
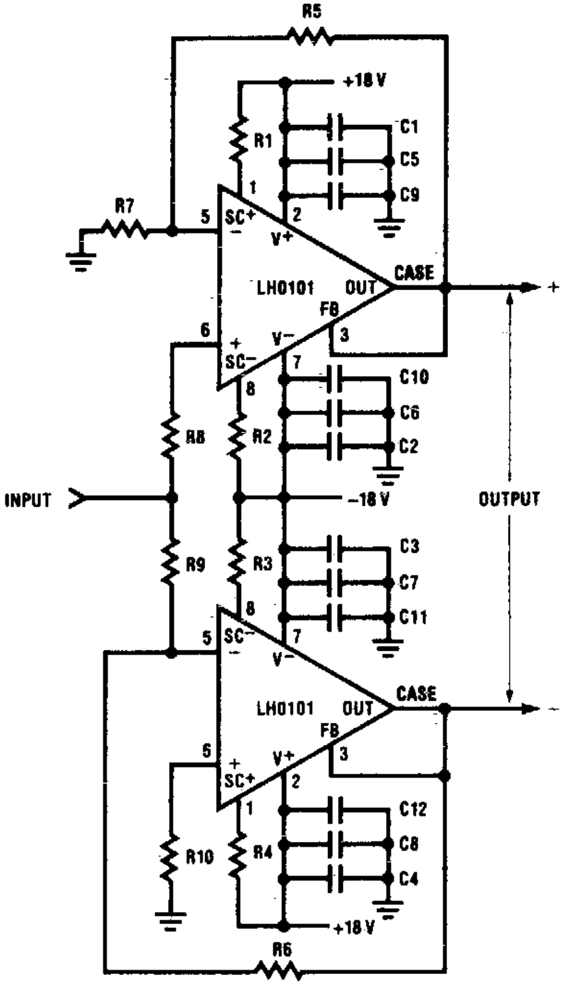
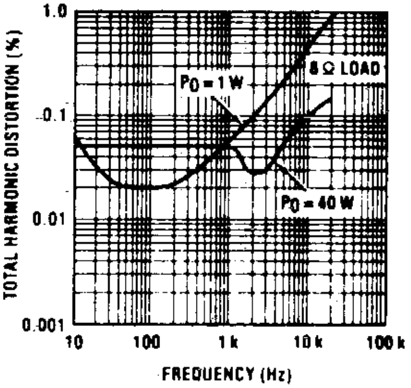

V_S
100k
0.1 μF
0.1 μF
100k
10k
1 μF
8
14
9
1
1/2 LM1877
13
3, 4, 5, 10, 11, 12
50 μF
500 μF
R_L
8Ω
2.7
0.1 μF

Fig. 8-20

BRIDGE AUDIO POWER AMPLIFIER


R5
+18 V
R1
C1
C5
C9
R7
5
SC+
1
2
V+
LH0101
OUT
CASE
FB
3
6
+
SC−
V−
7
8
C10
C6
C2
R8
R2
INPUT
−18 V
OUTPUT
+
−
R9
R3
C3
C7
C11
8
5
SC−
7
V−
LH0101
OUT
CASE
FB
6
+
V+
SC+
3
2
1
C12
C8
C4
R10
R4
+18 V
R6

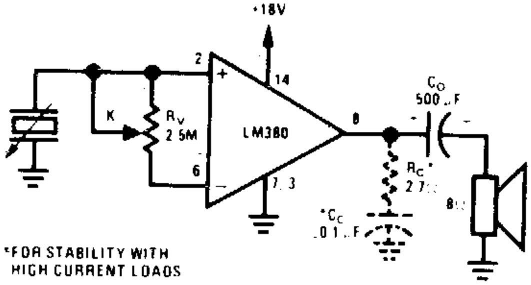


Total Harmonic Distortion vs. Frequency of Bridge Power Amplifier

R1-R4	CURRENT LIMIT RESISTOR	0.15 Ω 2 W
R5	FEEDBACK RESISTOR	5 kΩ
R6	FEEDBACK RESISTOR	15 kΩ
R7-R10	INPUT RESISTORS	10 kΩ
C1-C4	BYPASS CAPACITORS	47 µF 25 V ELECTROLYTIC
C5-C8	BYPASS CAPACITORS	10 µF 25 V TANTALUM
C9-C12	BYPASS CAPACITORS	0.1 µF 25V CERAMIC

Fig. 8-21

PHONO AMPLIFIER


+18V
2
+
14
K
R_V
2.5M
LM380
8
C_O
500 µF
6
−
7, 3
R_C*
2.7Ω
*C_C
.01 µF
8Ω

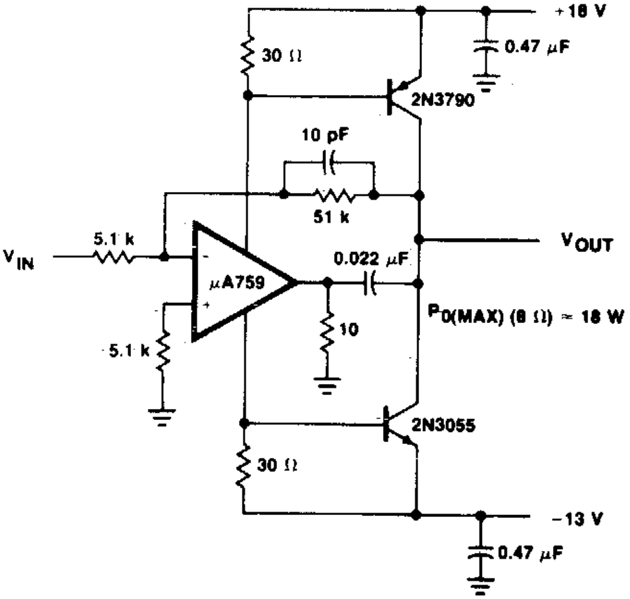
*FOR STABILITY WITH HIGH CURRENT LOADS

Circuit Notes

Used when maximum input impedance is required or the signal attenuation of the voltage divider volume control is undesirable.

Fig. 8-22

HIGH SLEW RATE POWER OP AMP/AUDIO AMP



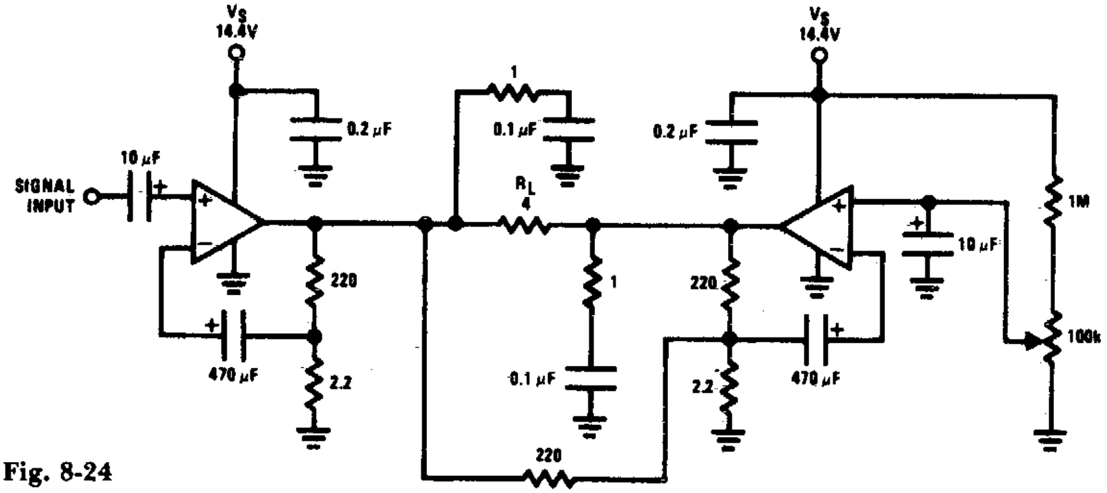
+18 V
30 Ω
0.47 μF
2N3790
10 pF
51 k
V_{IN}
5.1 k
μA759
0.022 μF
V_{OUT}
10
$P_{O(MAX)}$ (8 Ω) ≈ 18 W
5.1 k
2N3055
30 Ω
−13 V
0.47 μF

Fig. 8-23

Features

- High Slew Rate 9 V/μs
- High 3 dB Power Bandwidth 85 kHz
- 18 Watts Output Power Into an 8 Ω Load.
- Low Distortion — .2%, 10 VRMS, 1 kHz Into 8 Ω

16 W BRIDGE AMPLIFIER



V_S 14.4V
0.2 μF
10 μF
SIGNAL INPUT
220
470 μF
2.2
1
0.1 μF
R_L 4
1
0.1 μF
220
V_S 14.4V
0.2 μF
10 μF
1M
100k
220
2.2
470 μF

Fig. 8-24

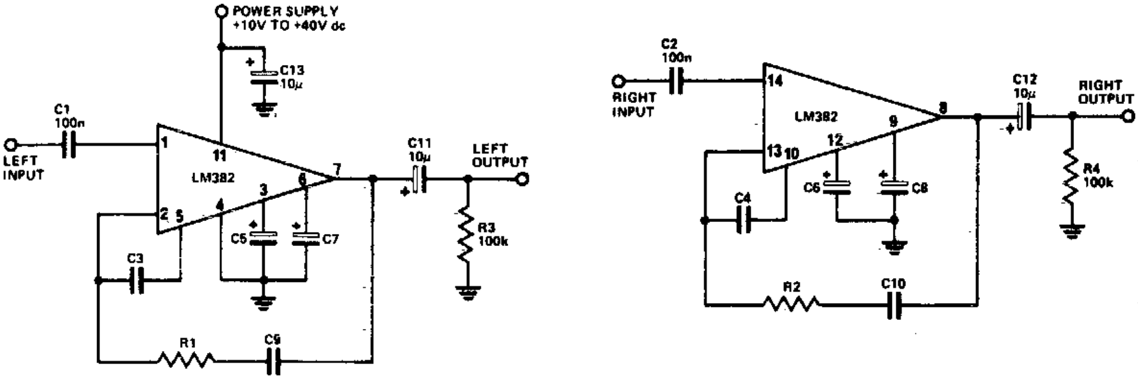
9

Audio Signal Amplifiers

The sources of the following circuits are contained in the Sources section beginning on page 730. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

General Purpose Preamplifier
Basic Transistor Amplifier Circuits
Microphone Amplifier
Transducer Amplifier
Ultra-High Gain Audio Amplifier
Transformerless Microphone Preamp (Balanced Inputs)
Transformerless Microphone Preamp (Unbalanced Inputs)
Magnetic Pickup Phone Preamplifier
Disc/Tape Phase Modulated Readback Systems
Two-Pole Fast Turn-On NAB Tape Preamplifier
Tape Preamplifier (NAB Equation)
LM382 Phono Preamplifier
Tape Recording Amplifier
Magnetic Phono Preamplifier
Phono Preamp
Remote Amplifier
Adjustable Gain Noninverting Amplifier
High Gain Inverting AC Amplifier
Flat Response Amplifier
Preamplifier with RIAA/NAB Compensation
Tape Playback Amplifier

GENERAL PURPOSE PREAMPLIFIER


POWER SUPPLY
+10V TO +40V dc
C13
10μ
C1
100n
LEFT
INPUT
LM382
1
11
2
5
3
4
6
7
C5
C7
C3
R1
C9
C11
10μ
LEFT
OUTPUT
R3
100k
C2
100n
RIGHT
INPUT
14
LM382
9
8
12
13
10
C4
C6
C8
R2
C10
C12
10μ
RIGHT
OUTPUT
R4
100k

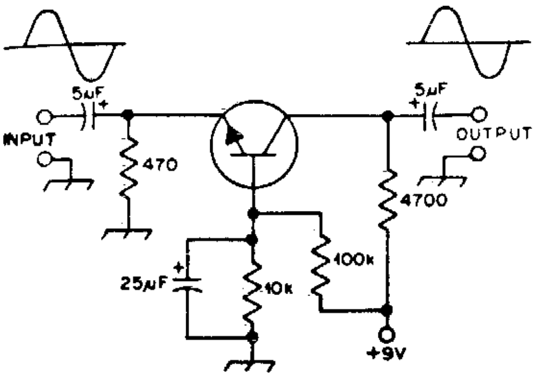
FUNCTION	C3, 4	C5, 6	C7, 8	C9, 10	R1, 2
Phono preamp (RIAA)	330n	10μF	10μF	1n5	1k
Tape preamp (NAB)	68n	10μF	10μF	–	–
Flat 40dB gain	–	–	10μF	–	–
Flat 55dB gain	–	10μF	–	–	–
Flat 80dB gain	–	10μF	10μF	–	–

Fig. 9-1

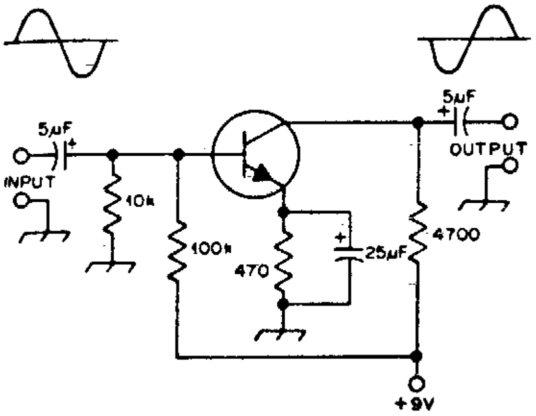
Circuit Notes

Not much can be said about how the LM382 works as most of the circuitry is contained within the IC. Most of the frequency-determining components are on the chip—only the capacitors are mounted externally. The LM382 has the convenient characteristic of rejecting ripple on the supply line by about 100 dB, thus greatly reducing the quality requirment for the power supply.

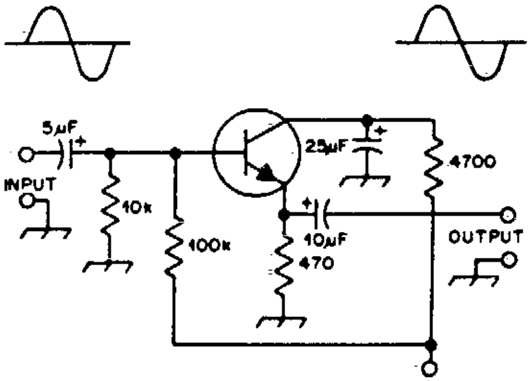
BASIC TRANSISTOR AMPLIFIER CIRCUITS


5µF
INPUT
470
5µF
OUTPUT
4700
100k
25µF
10k
+9V

COMMON BASE


5µF
OUTPUT
5µF
INPUT
10k
100k
470
25µF
4700
+9V

COMMON EMITTER


5µF
INPUT
25µF
4700
10k
100k
10µF
470
OUTPUT
+9V

COMMON COLLECTOR

Circuit Notes

Typical component values are given for use at audio frequencies, where these circuits are used most often. The input and output phase relationships are shown.

Fig. 9-2

ELECTRONIC BALANCED INPUT MICROPHONE AMPLIFIER

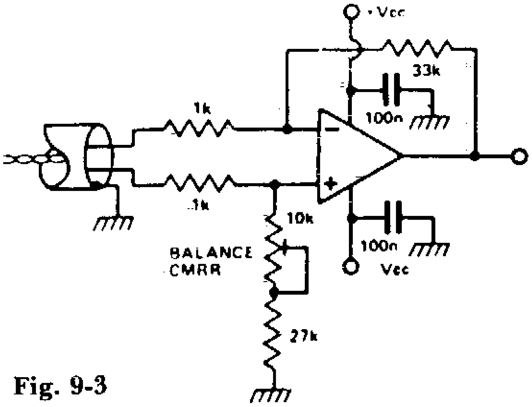

• Vcc
33k
100n
1k
1k
10k
BALANCE
CMRR
100n
Vcc
27k

Fig. 9-3

Circuit Notes

It is possible to simulate the balanced performance of a transformer electronically with a different amplifier. By adjusting the presets, the resistor ratio can be balanced so that the best CMRR is obtained. It is possible to get a better CMRR than from a transformer. Use a RC4136 which is a quad low noise op amp.

TRANSDUCER AMPLIFIER

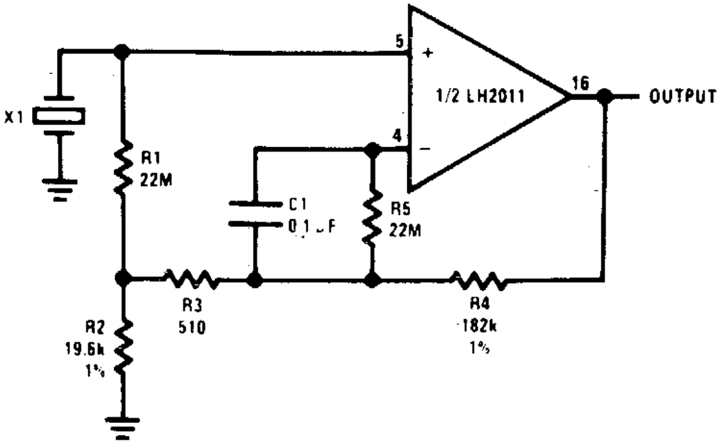

5
+
1/2 LH2011
16
OUTPUT
4
−
X1
R1
22M
C1
0.1 μF
R5
22M
R3
510
R4
182k
1%
R2
19.6k
1%

Fig. 9-4

$$R_{IN} = R1\left(1 + \frac{R2}{R3}\right) \quad A_V = \frac{R2 + R3 + R4}{R2 + R3}$$

Circuit Notes

This circuit is high-input-impedance ac amplifier for a piezoelectric transducer. Input resistance is 880 M, and a gain of 10 is obtained.

ULTRA-HIGH GAIN AUDIO AMPLIFIER

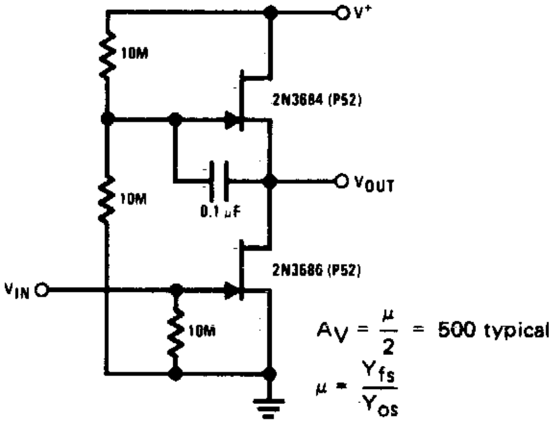

V+
10M
2N3684 (P52)
10M
0.1 μF
VOUT
2N3686 (P52)
VIN
10M
$A_V = \frac{\mu}{2} = 500$ typical
$\mu = \frac{Y_{fs}}{Y_{os}}$

Fig. 9-5

Circuit Notes

Sometimes called the JFET μ-amp, this circuit provides a very low power, high gain amplifying function. Since μ of a JFET increases as drain current decreases, the lower drain current is, the more gain you get. Input dynamic range is sacrificed with increasing gain, however.

MICROPHONE AMPLIFIER

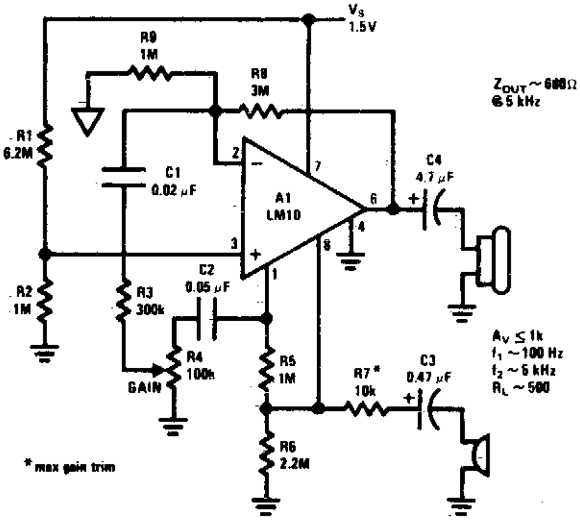
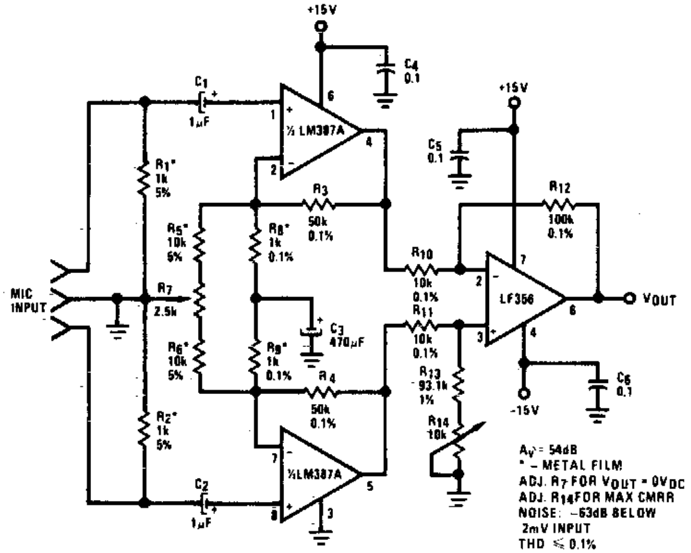

V_S
1.5V
R9
1M
R8
3M
$Z_{OUT} \sim 600\Omega$
@ 5 kHz
R1
6.2M
C1
0.02 μF
2
7
A1
LM10
6
4
8
3
1
C4
4.7 μF
C2
0.05 μF
R2
1M
R3
300k
R4
100k
GAIN
R5
1M
R7*
10k
C3
0.47 μF
$A_V \leq 1k$
$f_1 \sim 100$ Hz
$f_2 \sim 5$ kHz
$R_L \sim 500$
R6
2.2M
* max gain trim

Fig. 9-6

Circuit Notes

This circuit operates from a 1.5 Vdc source.

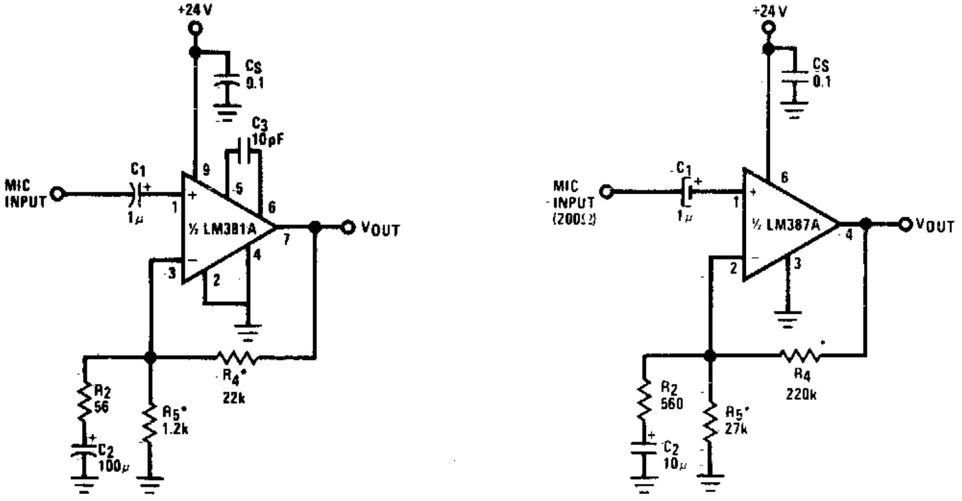
TRANSFORMERLESS (BALANCE INPUTS) MICROPHONE PREAMP


+15V
C4
0.1
C1
1μF
½ LM387A
1
2
4
6
+15V
C5
0.1
R1*
1k
5%
R3
50k
0.1%
R12
100k
0.1%
R5*
10k
5%
R8*
1k
0.1%
R10
10k
0.1%
MIC
INPUT
R7
2.5k
LF356
2
3
4
6
7
VOUT
C3
470μF
R11
10k
0.1%
R6*
10k
5%
R9*
1k
0.1%
R4
50k
0.1%
R13
93.1k
1%
C6
0.1
-15V
R2*
1k
5%
R14
10k
½LM387A
7
8
3
5
C2
1μF

$A_V = 54dB$
* – METAL FILM
ADJ. R_7 FOR $V_{OUT} = 0V_{DC}$
ADJ. R_{14} FOR MAX CMRR
NOISE: –63dB BELOW
2mV INPUT
THD ≤ 0.1%

Fig. 9-7

TRANSFORMERLESS MICROPHONE PREAMPS (UNBALANCED INPUTS)


+24V
CS
0.1
C3
10pF
MIC
INPUT
C1
1μ
½ LM381A
9
5
1
6
7
3
2
4
VOUT
R2
56
C2
100μ
R5*
1.2k
R4*
22k
+24V
CS
0.1
MIC
INPUT
(200Ω)
C1
1μ
½ LM387A
6
1
2
3
4
VOUT
R2
560
C2
10μ
R5*
27k
R4
220k

A_V = 52dB
* – METAL FILM
NOISE: -69dB BELOW
2mV
THD 0.1%

A_V = 52dB
* – METAL FILM
NOISE: -67dB BELOW
2mV
THD 0.1%

Fig. 9-8 (a) LM381A S. E. Bias (b) LM387A

MAGNETIC PICKUP PHONO PREAMPLIFIER

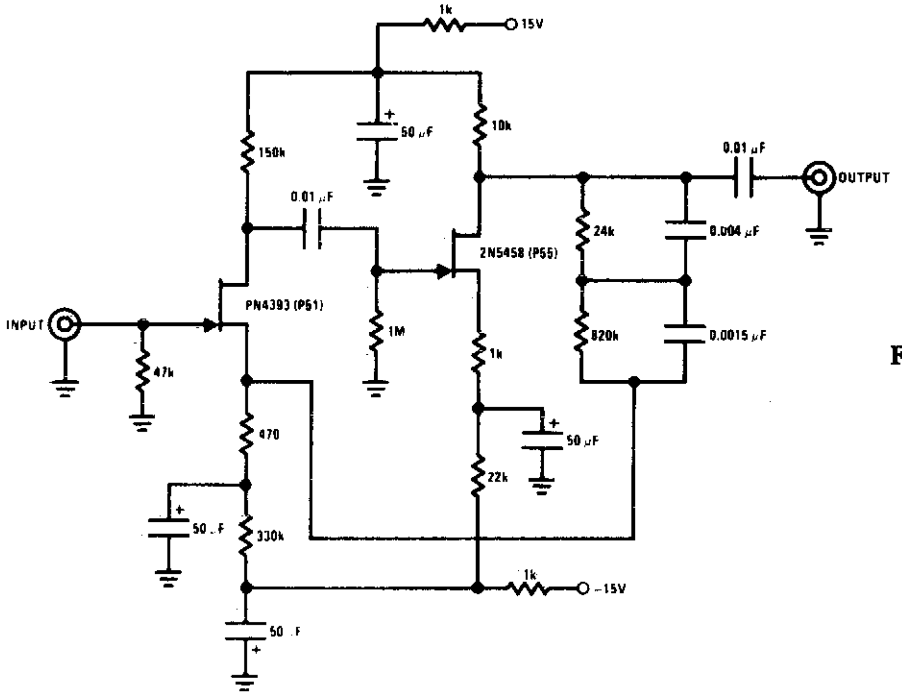

1k
15V
150k
50 μF
10k
0.01 μF
0.01 μF
OUTPUT
24k
0.004 μF
2N5458 (P55)
PN4393 (P51)
INPUT
1M
820k
0.0015 μF
47k
1k
50 μF
470
22k
50 μF
330k
1k
−15V
50 μF

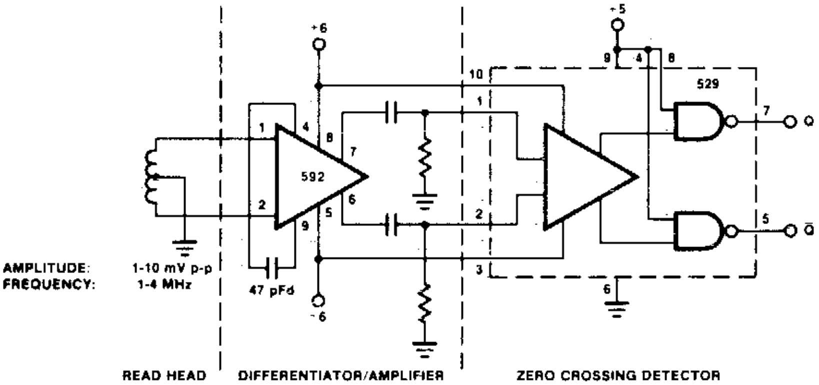
Fig. 9-9

Circuit Notes

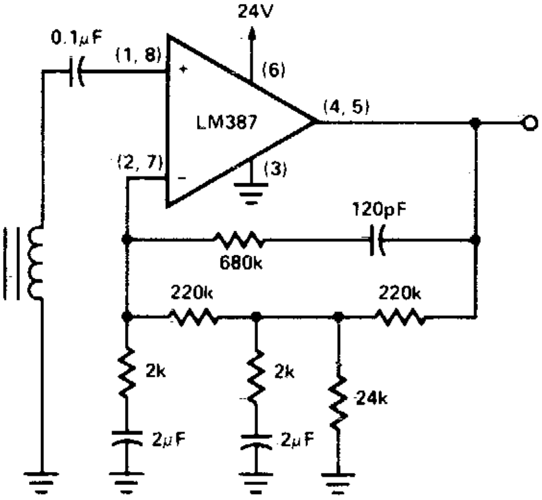
This preamplifier provides proper loading to a reluctance phono cartridge. It provides approximately 35 dB of gain at 1 kHz (2.2 mV input for 100 mV output). It features (S + N)/N ratio of better than −70 dB (referenced to 10 mV input at 1 kHz) and has a dynamic range of 84 dB (referenced to 1 kHz). The feedback provides for RIAA equalization.

DISC/TAPE PHASE MODULATED READBACK SYSTEMS

Fig. 9-10


+6
+5
9
4
8
10
529
1
4
8
7
Q
592
7
1
2
6
5
9
2
5
$\bar{Q}$
3
6
AMPLITUDE: 1-10 mV p-p
FREQUENCY: 1-4 MHz
47 pFd
−6
READ HEAD
DIFFERENTIATOR/AMPLIFIER
ZERO CROSSING DETECTOR

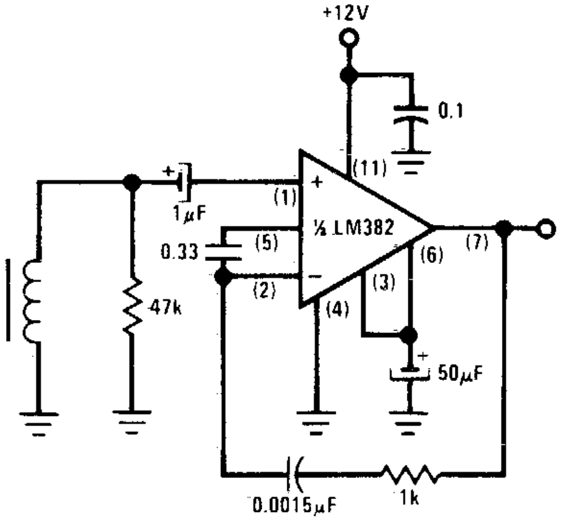
TWO-POLE FAST TURN-ON NAB TAPE PREAMPLIFIER



24V
0.1μF
(1, 8)
+
(6)
LM387
(4, 5)
(2, 7)
−
(3)
120pF
680k
220k
220k
2k
2k
24k
2μF
2μF

Fig. 9-11

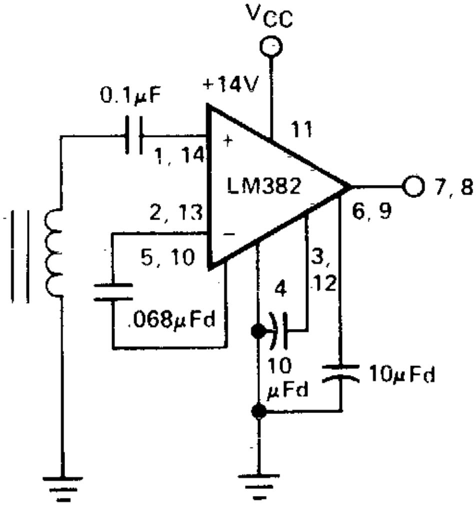
LM382 PHONO PREAMPLIFIER (RIAA)



+12V
0.1
(11)
+
1μF
(1)
+
½ LM382
0.33
(5)
(6)
(7)
−
(2)
(3)
(4)
47k
+
50μF
0.0015μF
1k

Fig. 9-13

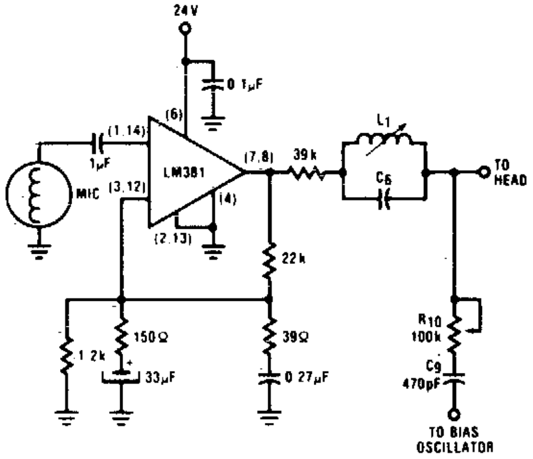
TAPE PREAMPLIFIER (NAB EQUALIZATION)



V_CC
+14V
0.1μF
11
1, 14
+
LM382
7, 8
2, 13
6, 9
−
5, 10
3,
12
4
.068μFd
10
μFd
10μFd

Fig. 9-12

TAPE RECORDING AMPLIFIER



24V
0.1μF
(6)
(1,14)
1μF
LM381
(7,8)
39k
L1
C6
TO HEAD
MIC
(3,12)
(4)
(2,13)
22k
150Ω
39Ω
1.2k
+
33μF
0.27μF
R10
100k
C9
470pF
TO BIAS OSCILLATOR

Fig. 9-14

MAGNETIC PHONO PREAMPLIFIER

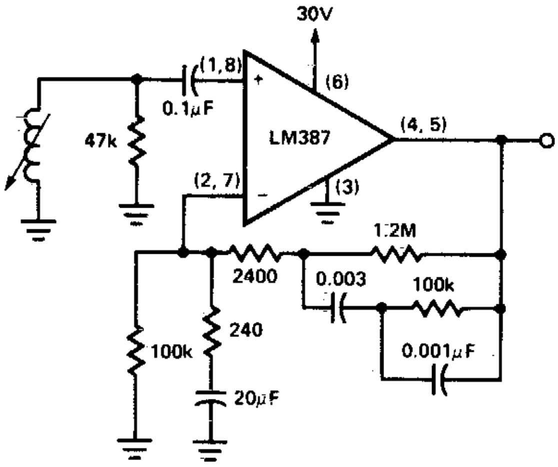

30V
(1,8)
(6)
0.1μF
47k
LM387
(4, 5)
(2, 7)
(3)
1.2M
2400
0.003
100k
240
100k
0.001μF
20μF

Fig. 9-15

REMOTE AMPLIFIER

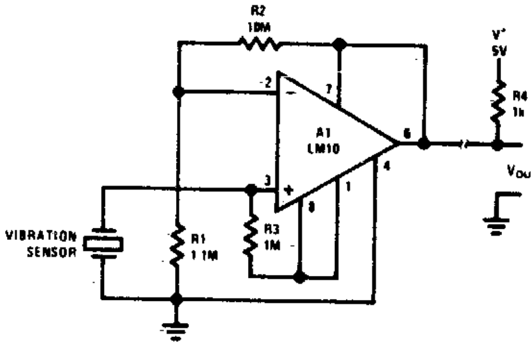
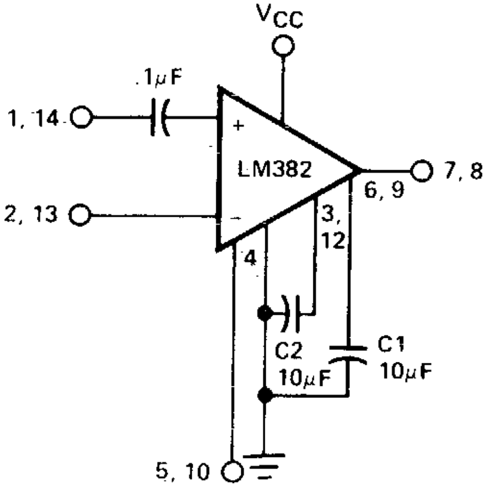

R2
10M
V+
5V
R4
1k
2
7
A1
LM10
6
4
3
8
1
VOUT
VIBRATION
SENSOR
R1
1.1M
R3
1M

Fig. 9-17

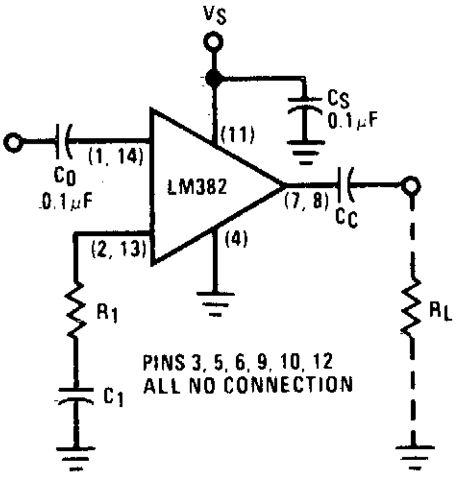
PHONO PREAMP (RIAA EQUALIZATION)


VCC
.1μF
1, 14
+
LM382
7, 8
6, 9
2, 13
3,
12
4
C2
10μF
C1
10μF
5, 10

CAPACITOR	GAIN
C1 Only	40dB
C2 Only	55dB
C1 & C2	80dB

Fig. 9-16

ADJUSTABLE GAIN NONINVERTING AMPLIFIER


VS
CS
0.1μF
(11)
(1, 14)
C0
0.1μF
LM382
(7, 8)
CC
(4)
(2, 13)
R1
RL
C1
PINS 3, 5, 6, 9, 10, 12
ALL NO CONNECTION

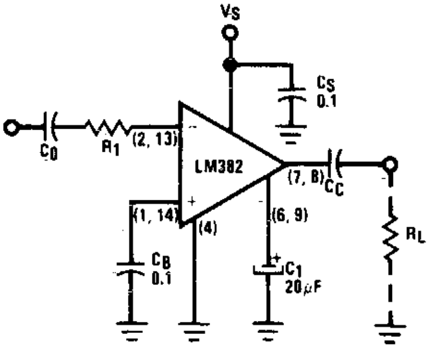
$$\text{GAIN} = 1 + \frac{267k}{R_1}$$

$$C_1 = \frac{1}{2\pi f_0 R_1}$$

f_0 = LOW FREQUENCY -3dB CORNER

Fig. 9-18

HIGH GAIN INVERTING AC AMPLIFIER


Vs
Cs
0.1
C0
R1
(2, 13)
LM382
(7, 8)
CC
(1, 14)
(4)
(6, 9)
RL
CB
0.1
C1
20μF

$\text{GAIN} = -\frac{5.1 \times 10^6}{R_1}$

$C_0 = \frac{1}{2\pi f_0 R_1}$

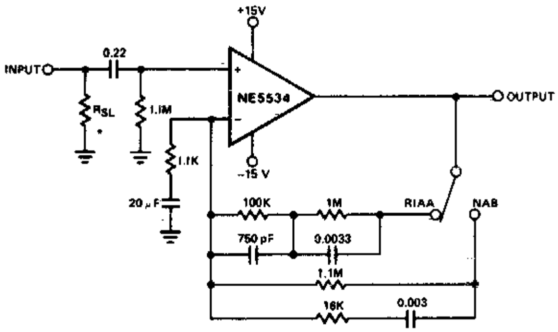
f_0 = LOW FREQUENCY -3dB CORNER $(C_c R_L >> C_0 R_1)$

INPUT IMPEDANCE = R_1

PINS 3, 5, 10, 12 NOT USED

Fig. 9-19

PREAMPLIFIER WITH RIAA/NAB COMPENSATION


+15V
0.22
INPUT
NE5534
OUTPUT
RSL
*
1.1M
1.1K
-15 V
20 μF
100K
1M
RIAA
NAB
750 pF
0.0033
1.1M
16K
0.003

*Select to provide specified transducer loading.
Output Noise ≅ 0.8mV rms (with input shorted)

All resistor values are in ohms.

Fig. 9-21

FLAT RESPONSE AMPLIFIER (FIXED GAIN CONFIGURATION)

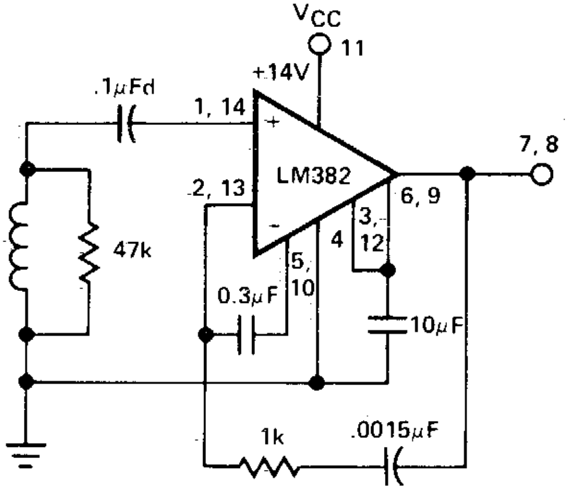

VCC
11
+14V
.1μFd
1, 14
LM382
7, 8
2, 13
6, 9
3, 12
4
5, 10
47k
0.3μF
10μF
1k
.0015μF

Fig. 9-20

TAPE PLAYBACK AMPLIFIER

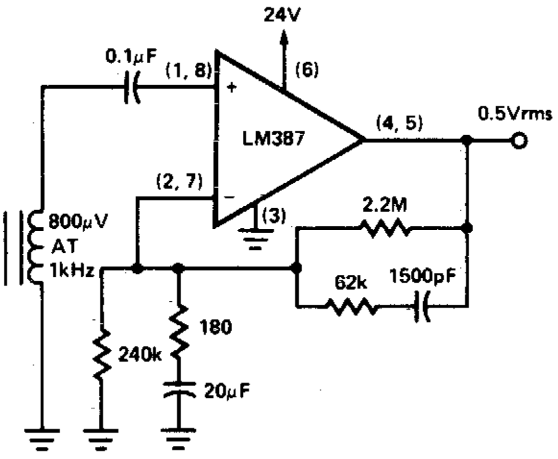

24V
0.1μF
(1, 8)
(6)
LM387
(4, 5)
0.5Vrms
(2, 7)
(3)
2.2M
800μV
AT
1kHz
62k
1500pF
180
240k
20μF

Fig. 9-22

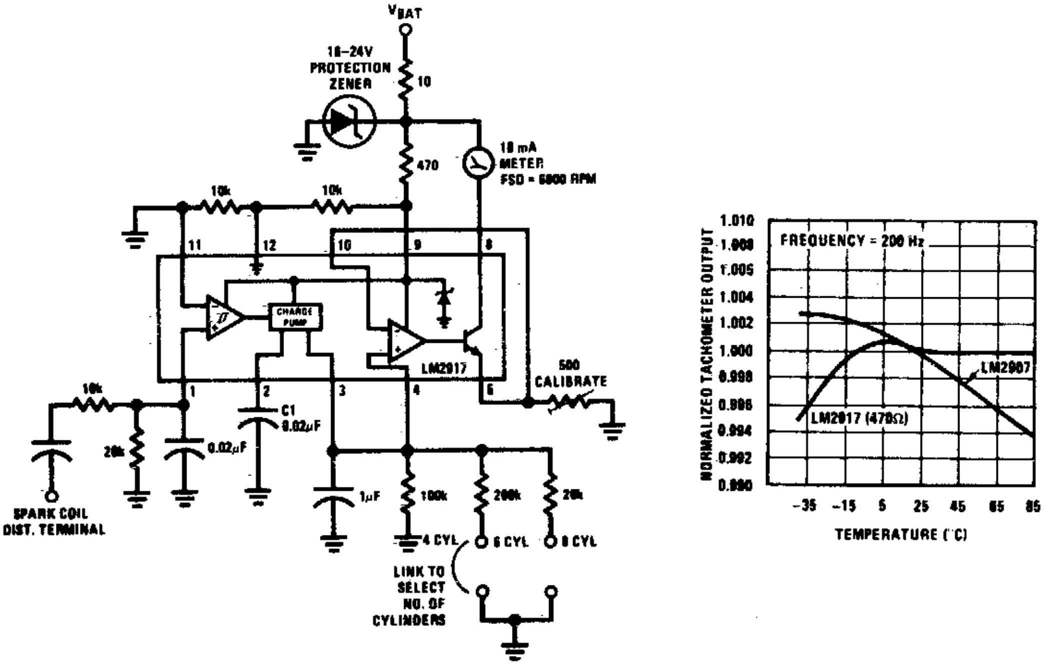
10

Automotive Circuits

The sources of the following circuits are contained in the Sources section beginning on page 730. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

Gasoline Engine Tachometer
Speed Alarm
Speed Warning Device
Universal Wiper Delay
Courtesy Light Extender
Bargraph Car Voltmeter
Tachometer
High Speed Warning Device
Breaker Point Dwell Meter
Tachometer
Capacitor Discharge Ignition System
Windshield Wiper Control
Auto Battery Current Analyzer
Speed Switch
Windshield Wiper Controller
Windshield Wiper Hesitation Control Unit
Ice Warning and Lights Reminder
Car Battery Monitor
Headlight Delay Unit
Windshield Washer Fluid Watcher
Car Battery Condition Checker
Overspeed Indicator
Sequential Flasher for Auto Turn Signals
Auto Lights-On Reminder

GASOLINE ENGINE TACHOMETER



V_BAT
18–24V PROTECTION ZENER
10
470
10 mA METER FSD = 6000 RPM
10k
10k
11
12
10
9
8
CHARGE PUMP
LM2917
500 CALIBRATE
1
2
3
4
5
10k
20k
0.02μF
C1 0.02μF
1μF
100k
200k
20k
SPARK COIL DIST. TERMINAL
4 CYL
6 CYL
8 CYL
LINK TO SELECT NO. OF CYLINDERS
NORMALIZED TACHOMETER OUTPUT
1.010
1.008
1.006
1.004
1.002
1.000
0.998
0.996
0.994
0.992
0.990
FREQUENCY = 200 Hz
LM2907
LM2917 (470Ω)
−35
−15
5
25
45
65
85
TEMPERATURE (°C)

Fig. 10-1

Circuit Notes

This tachometer can be set up for any number of cylinders by linking the appropriate timing resistor as illustrated. A 500 ohm trim resistor can be used to set up final calibration. A protection circuit composed of a 10 ohm resistor and a zener diode is also shown as a safety precaution against the transients which are to be found in automobiles.

SPEED ALARM

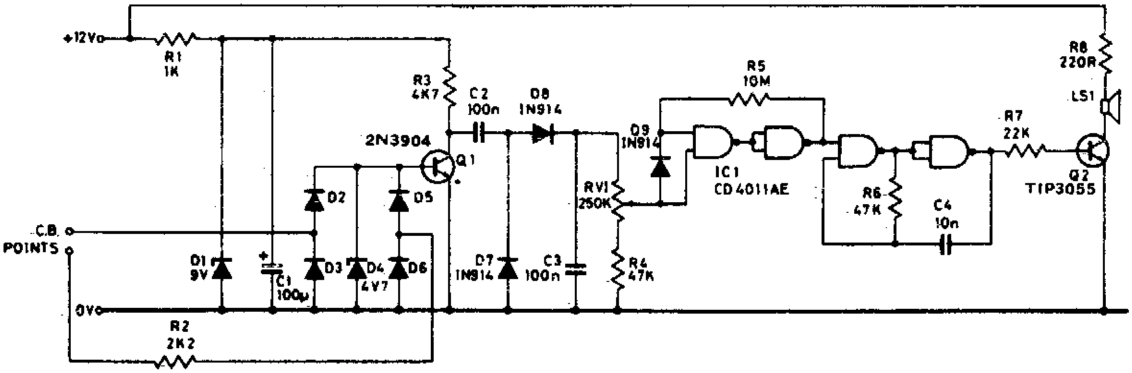
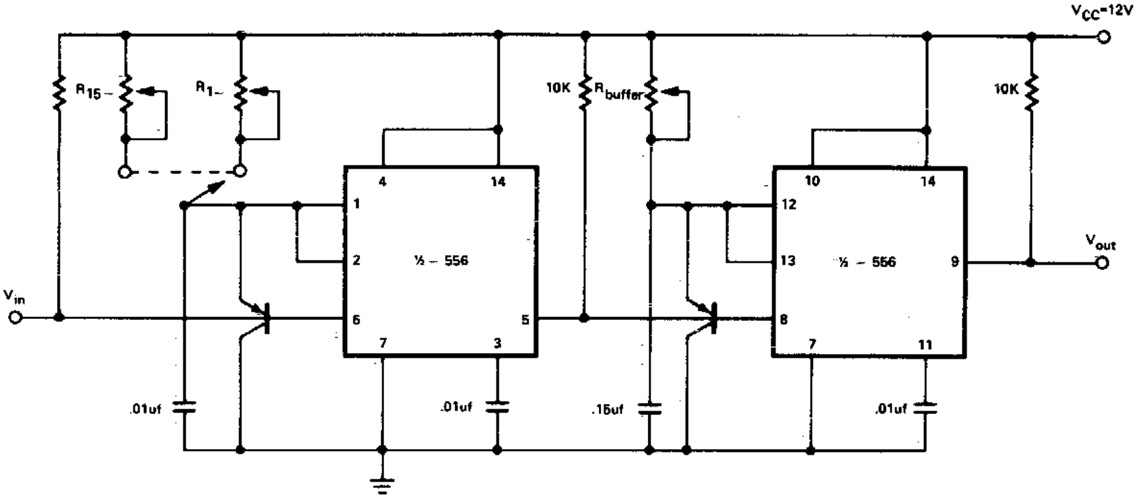

+12V
R1
1K
R3
4K7
C2
100n
D8
1N914
R5
10M
R8
220R
LS1
R7
22K
2N3904
Q1
D9
1N914
IC1
CD4011AE
Q2
TIP3055
D2
D5
RV1
250K
R6
47K
C4
10n
C.B.
POINTS
D1
9V
C1
100µ
D3
D4
4V7
D6
D7
1N914
C3
100n
R4
47K
0V
R2
2K2

Fig. 10-2

Circuit Notes

Pulses from the distributor points are passed through a current limiting resistor, rectified, and clipped at 4.7 volts. Via Q1 and the diode pump, a dc voltage proportional to engine rpm is presented to RV1; the sharp transfer characteristic of a CMOS gate, assisted by feedback, is used to enable the oscillator formed by the remaining half of the 4011. At the pre-set speed, a nonignorable tone emits from the speaker, and disappears as soon as the speed drops by three or four mph.

SPEED WARNING DEVICE


VCC=12V
R15
R1
10K
Rbuffer
10K
Vin
Vout
4
14
1
2
6
7
3
5
½ – 556
10
14
12
13
8
7
11
9
½ – 556
.01uf
.01uf
.15uf
.01uf

OPERATING WAVEFORMS

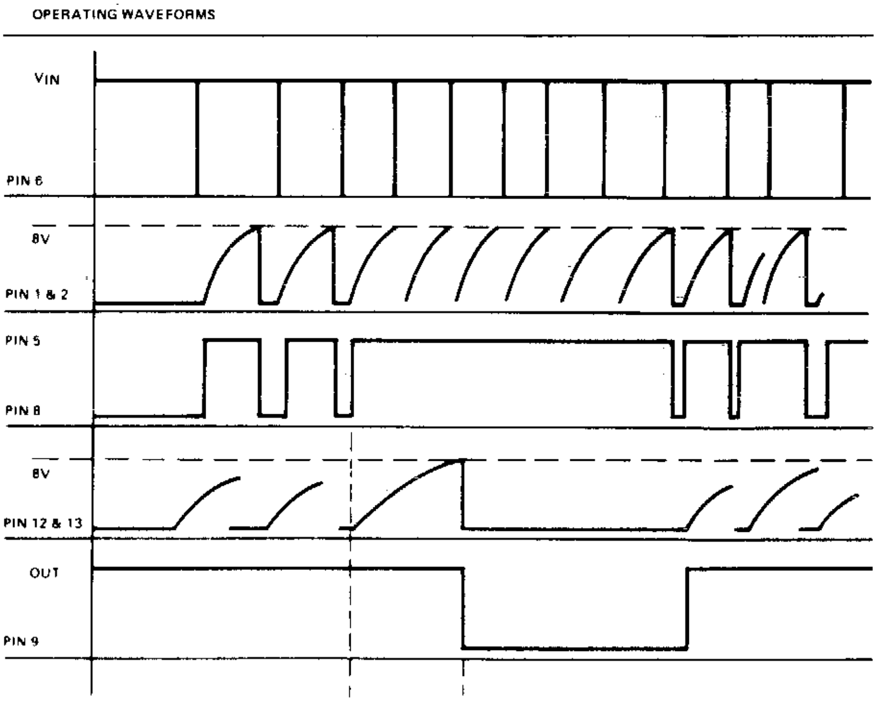


Fig. 10-3

UNIVERSAL WIPER DELAY

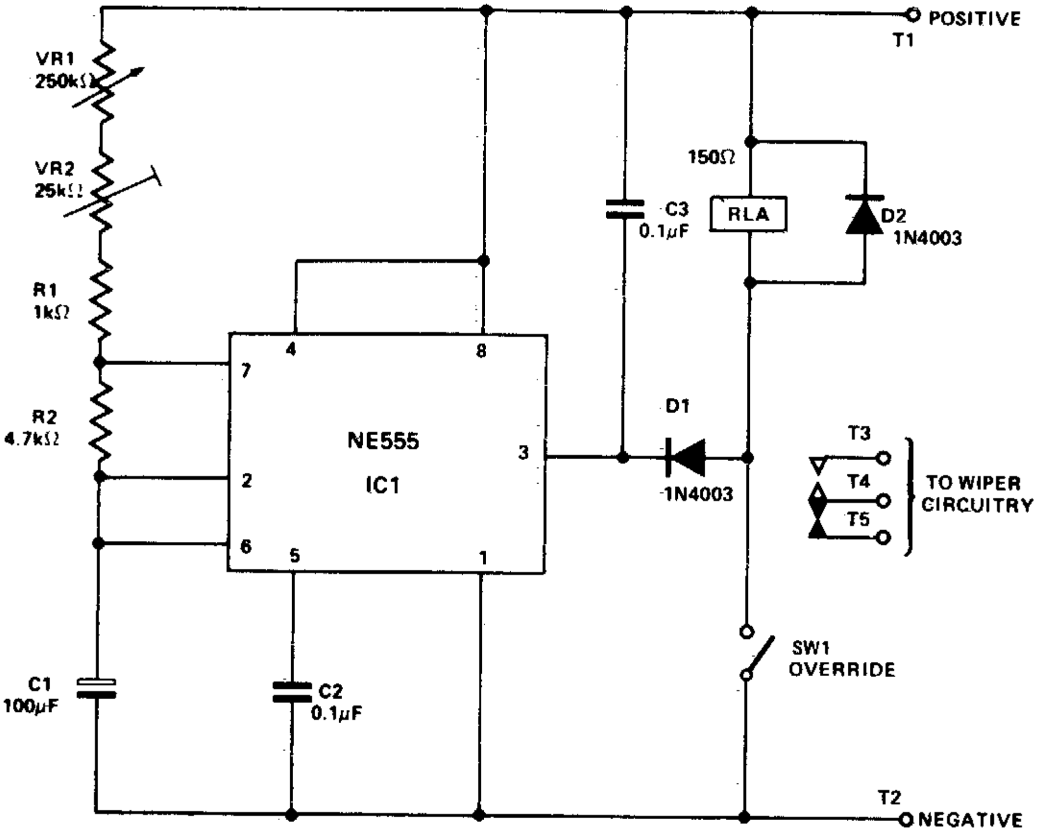

VR1
250kΩ
VR2
25kΩ
R1
1kΩ
R2
4.7kΩ
C1
100μF
C2
0.1μF
NE555
IC1
4
8
7
2
6
5
1
3
C3
0.1μF
150Ω
RLA
D2
1N4003
D1
1N4003
POSITIVE
T1
T3
T4
T5
TO WIPER CIRCUITRY
SW1
OVERRIDE
T2
NEGATIVE

Fig. 10-4

Circuit Notes

IC1 is connected in the astable mode, driving RLA. C3, D1, and D2 prevent spikes from the relay coil and the wiper motor from triggering IC1. VR2 is adjusted to give the minimum delay time required. VR1 is the main delay control and provides a range of from about 1 second to 20 seconds. SW1 is an override switch to hold RLA permanently on (for normal wiper operation). The relay should have a resistance of at least 150 ohms and have heavy duty contacts. The suppression circuit may be needed for the protection of IC1.

COURTESY LIGHT EXTENDER

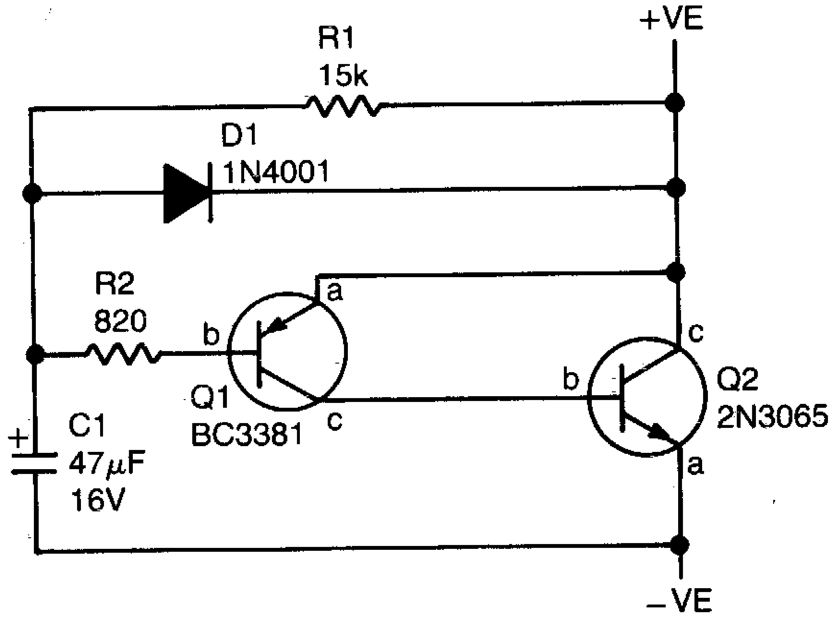


Fig. 10-5

Circuit Notes

Most car door switches are simply single-pole switches, with one side grounded. When the door is opened the switch grounds the other line thus completing the light circuit. In a car where the negative terminal of the battery is connected to the chassis, the negative wire of the unit (emitter of Q2) is connected to chassis the positive wire (case of 2N3055) is connected to the wire going to the switch. In a car having a positive ground system this connection sequence is reversed. When the switch closes (door open), C1 is discharged via D1 to zero volts, and when the switch opens, C1 charges up via R1 and R2. Transistors Q1 and Q2 are connected as an emitter follower (Q2 just buffers Q1) therefore the voltage across Q2 increases slowly as C1 charges. Hence Q2 acts like a low resistance in parallel with the switch and keeps the lights on. The value of C1 is chosen such that a useful light level is obtained for about four seconds; therefore the light decreases until in about 10 seconds it is out completely. With different transistor gains and with variation in current drain due to a particular type of car, the timing may vary but may be simply adjusted by selecting C1.

BARGRAPH CAR VOLTMETER

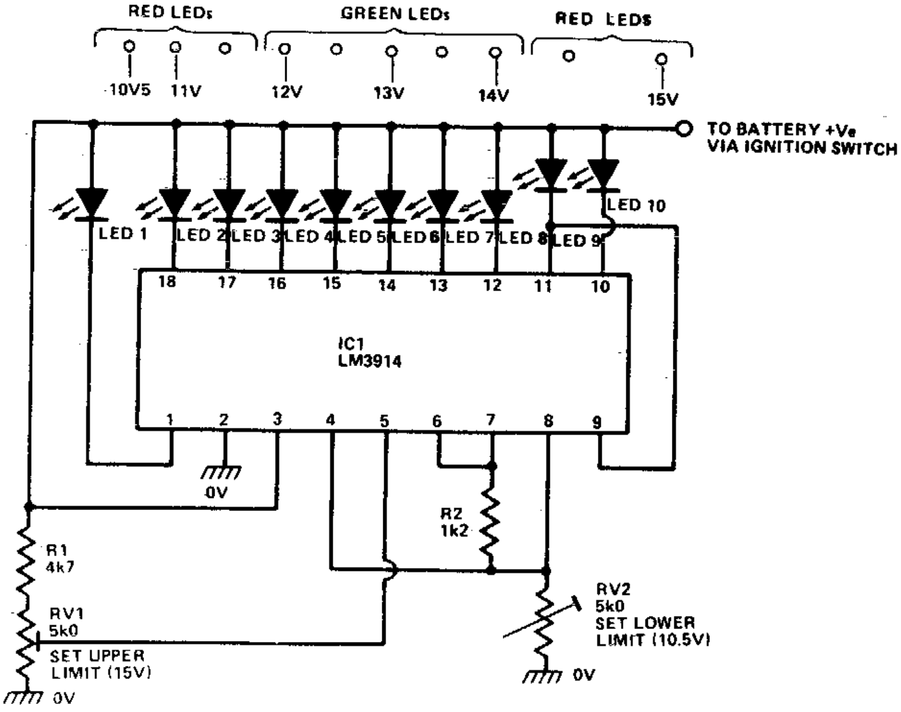

RED LEDs
GREEN LEDs
RED LEDS
10V5
11V
12V
13V
14V
15V
TO BATTERY +Ve VIA IGNITION SWITCH
LED 1
LED 2
LED 3
LED 4
LED 5
LED 6
LED 7
LED 8
LED 9
LED 10
18
17
16
15
14
13
12
11
10
IC1
LM3914
1
2
3
4
5
6
7
8
9
0V
R2
1k2
R1
4k7
RV1
5k0
SET UPPER LIMIT (15V)
0V
RV2
5k0
SET LOWER LIMIT (10.5V)
0V

Fig. 10-6

Circuit Notes

The LM3914 acts as a LED-driving voltometer that has its basic maximum and minimum readings determined by the values of R2 and RV2. When correctly adjusted, the unit actually covers the 2.5 volt to 3.6 volt range, but it is made to read a supply voltage span of 10-10.5 volts to 15 volts by interposing potential divider R1-RV1 between the supply line and the pin-5 input terminal of the IC. The IC is configured to give a 'dot' display, in which only one of the ten LEDs is illuminated at any given time. If the supply voltage is below 10.5 volts none of the LEDs illuminate. If the supply equals or exceeds 15 volts, LED 10 illuminates.

TACHOMETER

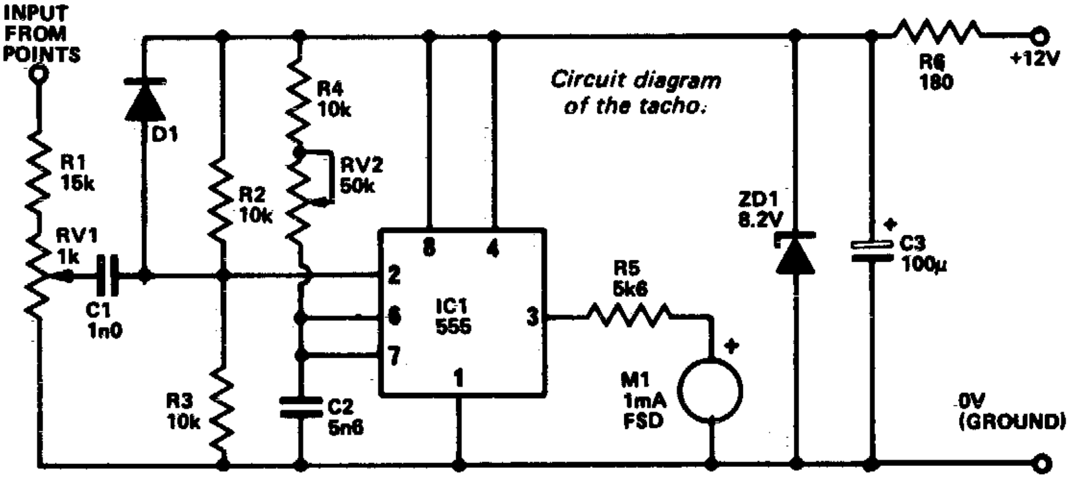

INPUT FROM POINTS
D1
R4 10k
RV2 50k
R1 15k
R2 10k
RV1 1k
C1 1n0
R3 10k
C2 5n6
8
4
2
6
7
IC1 555
3
1
Circuit diagram of the tacho.
R6 180
+12V
ZD1 8.2V
C3 100μ
R5 5k6
M1 1mA FSD
+
0V (GROUND)

Fig. 10-7

Circuit Notes

An electrical signal taken from the low tension side of the distributor is converted into a voltage proportional to engine rpm and this voltage is displayed on a meter calibrated accordingly. The 555 timer IC is used as a monostable which, in effect, converts the signal pulse from the breaker points to a single positive pulse the width of which is determined by the value of R4 + RV2 and C2. Resistors R2 and R3 set a voltage of about 4 volts at pin 2 of IC1. The IC is triggered if this voltage is reduced to less than approximately 2.7 volts (⅓ of supply voltage), and this occurs due to the voltage swing when the breaker points open. An adjustment potentiometer RV1 enables the input level to be set to avoid false triggering. Zener diode ZD1 and the 180 ohm resistor stabilize the unit against voltage variations.

HIGH SPEED WARNING DEVICE

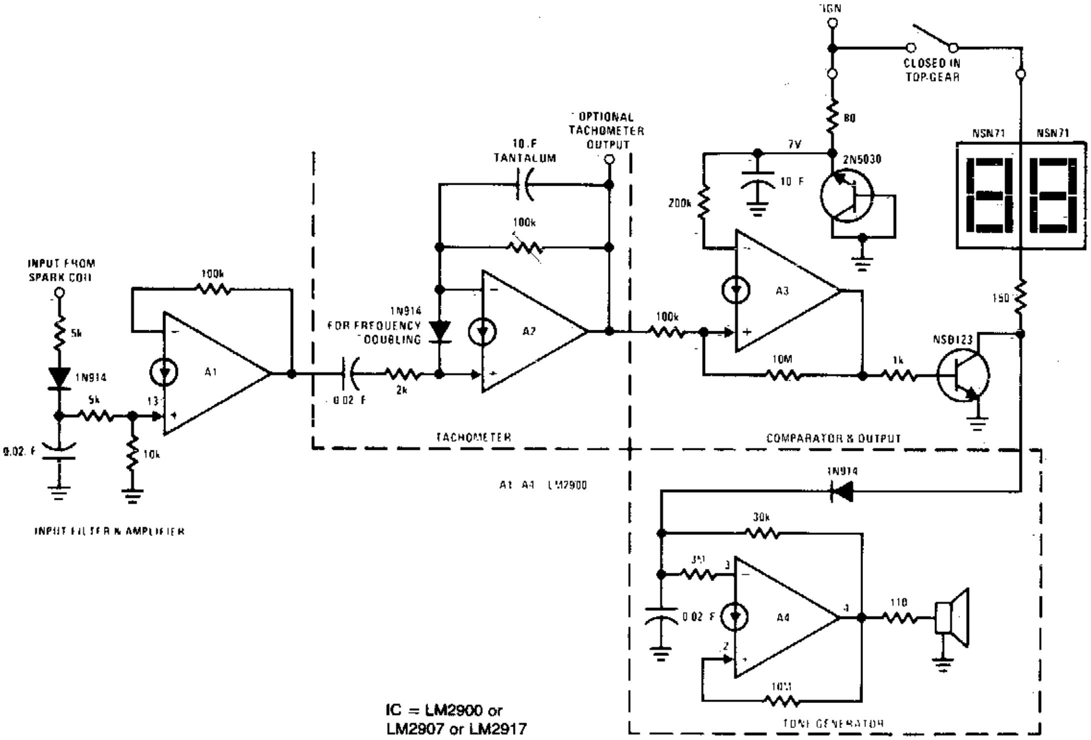

IGN
CLOSED IN TOP-GEAR
80
7V
2N5030
10 F
200k
NSN71
NSN71
OPTIONAL TACHOMETER OUTPUT
10 F TANTALUM
100k
INPUT FROM SPARK COIL
100k
5k
1N914
5k
13
A1
0.02 F
10k
INPUT FILTER & AMPLIFIER
1N914 FOR FREQUENCY DOUBLING
0.02 F
2k
A2
TACHOMETER
A1 A4 LM2900
100k
A3
10M
1k
150
NSB123
COMPARATOR & OUTPUT
1N914
30k
3M
3
0.02 F
A4
4
2
110
10M
TONE GENERATOR
IC = LM2900 or LM2907 or LM2917

Fig. 10-8

Circuit Notes

A1 amplifies and regulates the signal from the spark coil. A2 converts frequency to voltage so that its output is a voltage proportional to engine rpm. A3 compares the tachometer voltage with the reference voltage and turns on the output transistor at the set speed. Amplifier A4 is used to generate an audible tone whenever the set speed is exceeded.

BREAKER POINT DWELL METER

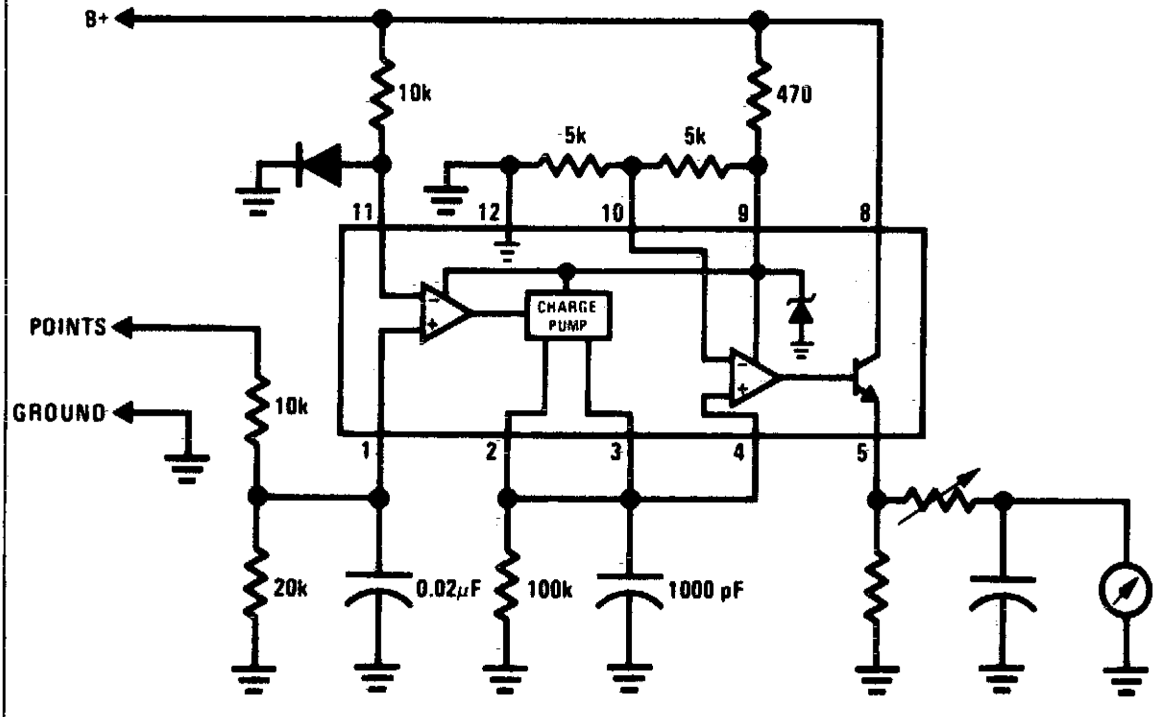
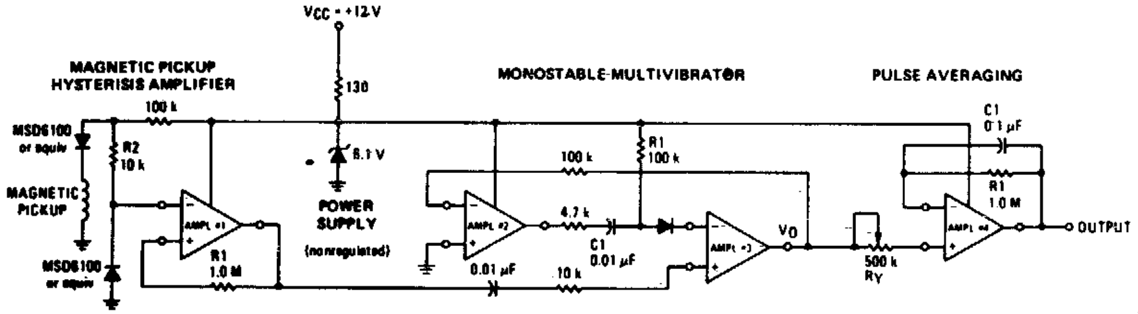

B+
10k
470
5k
5k
11
12
10
9
8
CHARGE PUMP
POINTS
GROUND
10k
1
2
3
4
5
20k
0.02μF
100k
1000 pF

Fig. 10-9

TACHOMETER


VCC = +12 V
MAGNETIC PICKUP HYSTERISIS AMPLIFIER
130
MONOSTABLE-MULTIVIBRATOR
PULSE AVERAGING
100 k
MSD6100 or equiv
R2 10 k
8.1 V
R1 100 k
100 k
C1 0.1 μF
MAGNETIC PICKUP
POWER SUPPLY
(nonregulated)
AMPL #1
AMPL #2
AMPL #3
AMPL #4
4.7 k
C1 0.01 μF
R1 1.0 M
V0
OUTPUT
MSD6100 or equiv
R1 1.0 M
0.01 μF
10 k
500 k RY

Hysterisis Voltage for Switching

$$V_H = \frac{A_I R2}{R1} \quad (V_{CC} - 1.6)$$

Timing-Interval: $t \approx 0.7\ R1\ C1$

$$V_{p\text{-}p} \approx \frac{(V_0 - 0.6) \cdot A_I \cdot t}{R_Y C1}$$

Fig. 10-10

CAPACITOR DISCHARGE IGNITION SYSTEM

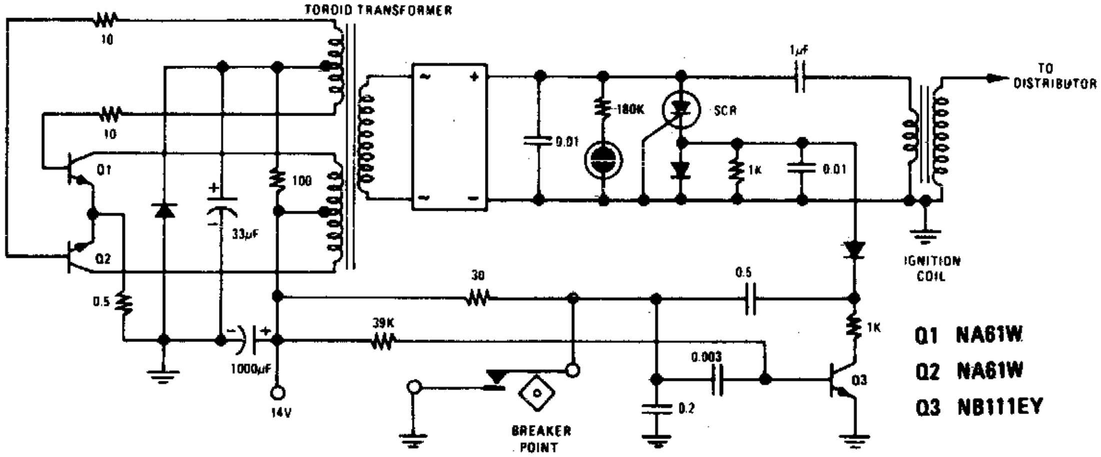

TOROID TRANSFORMER
10
10
Q1
Q2
0.5
33μF
100
1000μF
14V
0.01
180K
SCR
1K
0.01
1μF
TO DISTRIBUTOR
IGNITION COIL
30
39K
0.5
1K
0.003
0.2
Q3
BREAKER POINT
Q1 NA61W
Q2 NA61W
Q3 NB111EY

Fig. 10-11

WINDSHIELD WIPER CONTROL

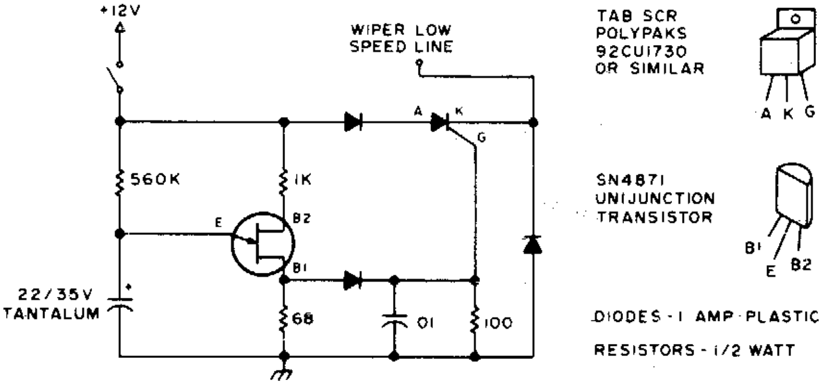

+12V
WIPER LOW SPEED LINE
A
K
G
560K
1K
E
B2
B1
22/35V TANTALUM
68
.01
100
TAB SCR POLYPAKS 92CU1730 OR SIMILAR
A K G
SN4871 UNIJUNCTION TRANSISTOR
B1
E
B2
DIODES - 1 AMP PLASTIC
RESISTORS - 1/2 WATT

Fig. 10-12

Circuit Notes

Here's a good way to set windshield wipers on an interval circuit. Only two connections to the car's wiper control, plus ground, are required. Variable control can be accomplished by substituting a 500 K pot in series with a 100 K fixed resistor in place of the 560 K.

AUTO BATTERY CURRENT ANALYZER

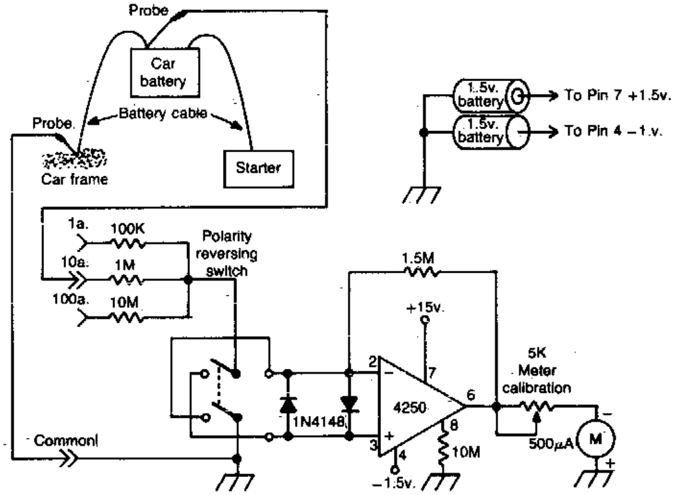
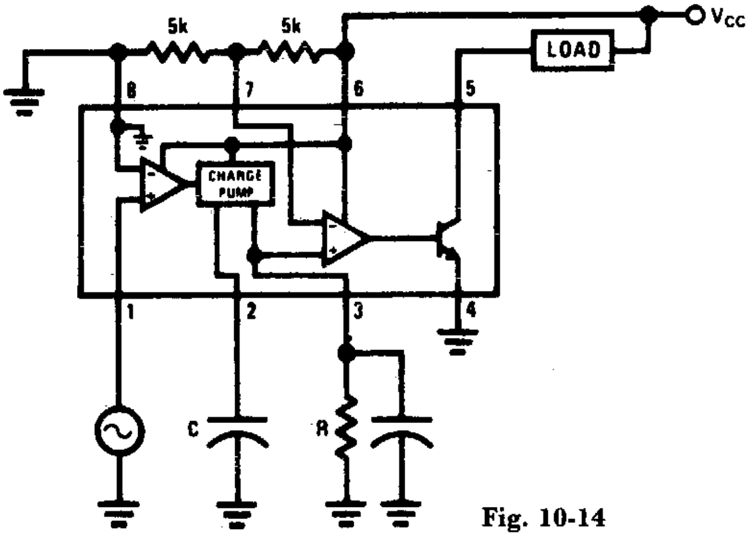

Probe
Car battery
Battery cable
Probe
Car frame
Starter
1.5v. battery
1.5v. battery
To Pin 7 +1.5v.
To Pin 4 −1.v.
1a.
100K
10a.
1M
100a.
10M
Polarity reversing switch
1.5M
+15v.
2
7
6
4250
1N4148
5K Meter calibration
Common
3
4
8
10M
−1.5v.
500µA
M

Fig. 10-13

Circuit Notes

This op-amp analyzer can measure the current drawn by any device in a car. The analyzer works by measuring the very small voltage that develops across the battery cables when current flows. To calibrate the unit, measure the current flow somewhere in the car with an accurate ammeter, then adjust the analyzer for that current reading.

SPEED SWITCH


5k
5k
Vcc
LOAD
8
7
6
5
CHARGE PUMP
1
2
3
4
C
R

Circuit Notes

Load is energized when

$$f_{in} \geqslant \frac{1}{2RC}.$$

Fig. 10-14

WINDSHIELD WIPER CONTROLLER

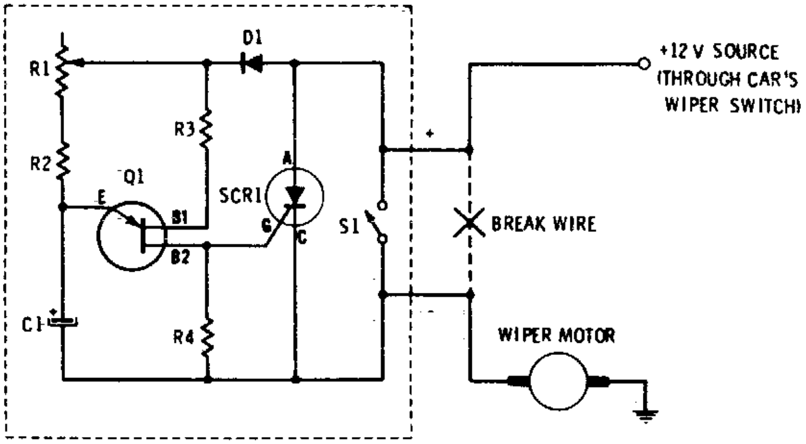

R1
D1
+12 V SOURCE
(THROUGH CAR'S
WIPER SWITCH)
R3
+
R2
Q1
E
A
SCR1
B1
G
C
S1
BREAK WIRE
B2
C1
R4
-
WIPER MOTOR

Fig. 10-15

Circuit Notes

This circuit provides complete speed control over car's windshield wipers. They can be slowed down to any rate even down to four sweeps per minute. The controller has two principal circuits: The rate-determining circuit—a unijunction transistor connected as a freerunning oscillator, and the silicon-controlled rectifier which is the actuator.

WINDSHIELD WIPER HESITATION CONTROL UNIT

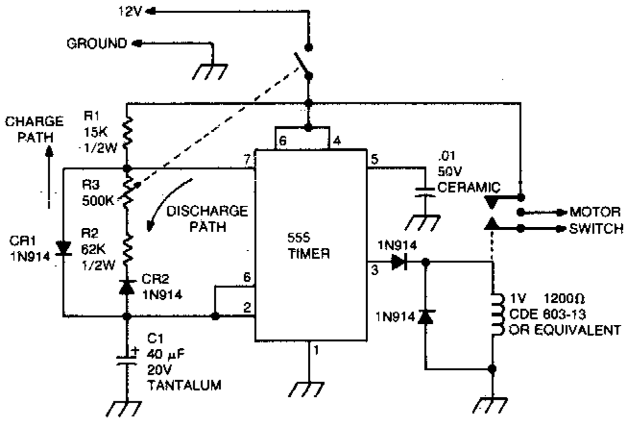
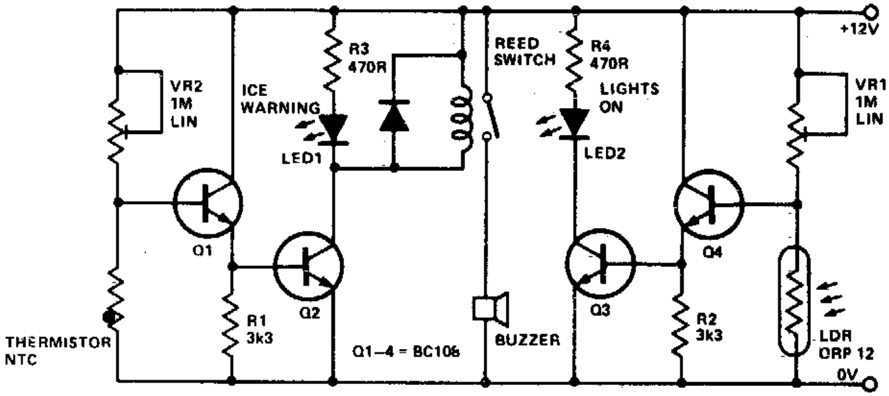

12V
GROUND
CHARGE PATH
R1
15K
1/2W
7
6
4
5
.01
50V
CERAMIC
R3
500K
DISCHARGE PATH
MOTOR
SWITCH
CR1
1N914
R2
62K
1/2W
555
TIMER
1N914
3
CR2
1N914
6
2
1V 1200Ω
CDE 803-13
OR EQUIVALENT
1N914
C1
40 µF
20V
TANTALUM
1

Fig. 10-16

Circuit Notes

This circuit uses the 555 timer in the astable or oscillatory mode. The length of time the timer is off is a function of the values of C1, R2, and R3. The potentiometer which controls the amount of "hesitation". (Approximately 2 to 15 seconds.) R2 provides a minimum time delay when R3 is at its zero ohms position.

ICE WARNING AND LIGHTS REMINDER



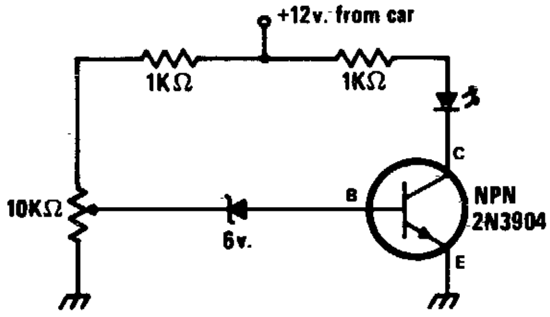
+12V
VR2
1M
LIN
ICE
WARNING
R3
470R
LED1
REED
SWITCH
R4
470R
LIGHTS
ON
LED2
VR1
1M
LIN
Q1
Q2
Q3
Q4
R1
3k3
R2
3k3
THERMISTOR
NTC
Q1–4 = BC106
BUZZER
LDR
ORP 12
0V

Fig. 10-17

Circuit Notes

This device will tell a driver if his lights should be on and will warn him if the outside temperature is nearing zero by lighting a LED and sounding a buzzer9 VR1 adjusts sensitivity for temperature, VR2 for light. Both thermistor and LDR should be well protected. Most high gain NPN transistors will work.

CAR BATTERY MONITOR



+12v. from car
1KΩ
1KΩ
C
B
E
NPN
2N3904
10KΩ
6v.

Circuit Notes

Warning light (LED) indicates when battery voltage falls below level set by 10 K pot. Can indicate that battery is defective or needs charging if cranking drops battery voltage below preset "safe" limit.

Fig. 10-18

HEADLIGHT DELAY UNIT

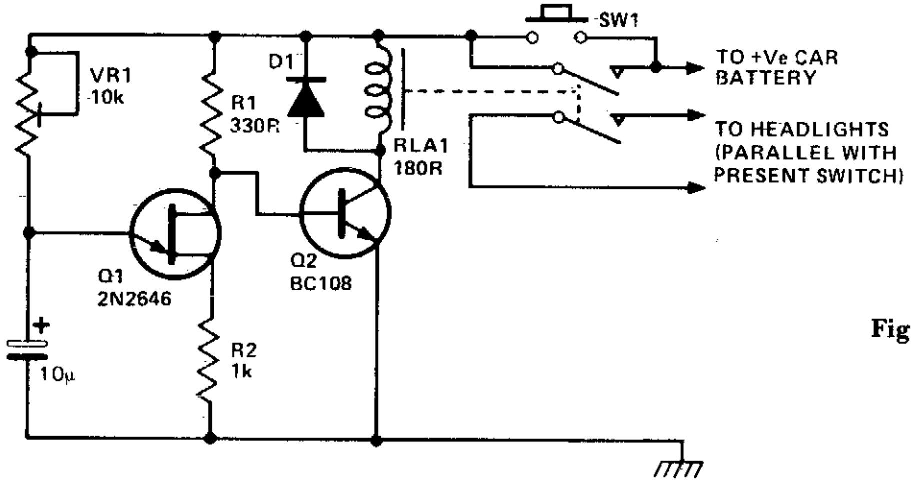

SW1
TO +Ve CAR BATTERY
TO HEADLIGHTS (PARALLEL WITH PRESENT SWITCH)
VR1 10k
D1
R1 330R
RLA1 180R
Q1 2N2646
Q2 BC108
+
10µ
R2 1k

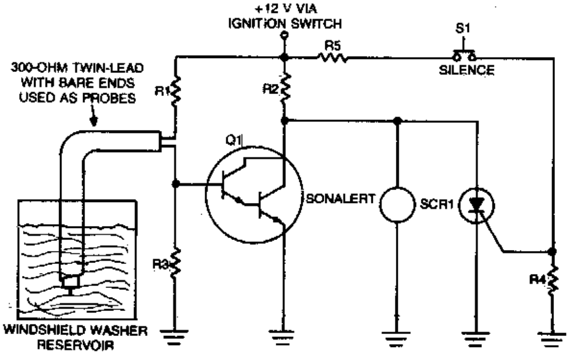
Fig. 10-19

Circuit Notes

This circuit will operate a car's headlights for a predetermined time to light up the driveway or path after the driver has left the car. SQ1 is pushed and Q2 is turned on closing the relay and turning on the car's headlights. C1 begins to charge through VR1 until Q1 turns on, turning Q2 off. The relay will then open switching off both the lights and the unit. The delay is governed by the time taken for the capacitor to charge, which is about one minute.

WINDSHIELD WASHER FLUID WATCHER

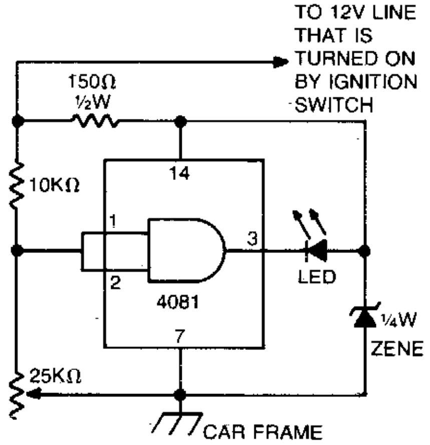
Fig. 10-20


+12 V VIA IGNITION SWITCH
S1
R5
SILENCE
300-OHM TWIN-LEAD WITH BARE ENDS USED AS PROBES
R1
R2
Q1
SONALERT
SCR1
R3
R4
WINDSHIELD WASHER RESERVOIR

Circuit Notes

This circuit relies upon the minute current between two conductive probes suspended in a washer fluid reservoir. When the level is below the probes, Q1 turns on and the Sonolert sounds.

CAR BATTERY CONDITION CHECKER

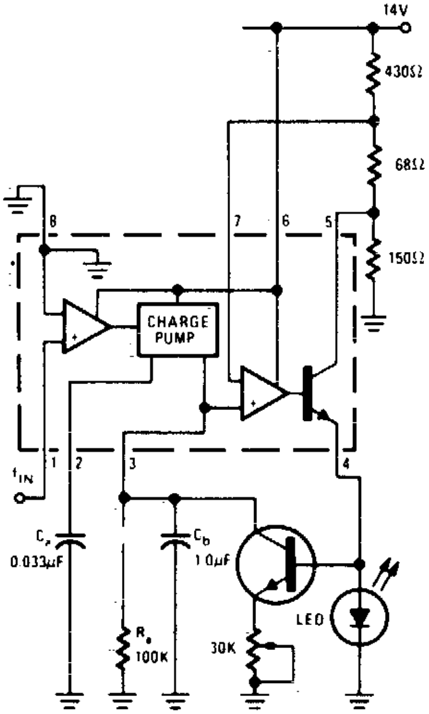

TO 12V LINE THAT IS TURNED ON BY IGNITION SWITCH
150Ω
½W
10KΩ
14
1
3
2
4081
7
LED
¼W
ZENE
25KΩ
CAR FRAME

Circuit Notes

This circuit uses an LED and 4081 CMOS integrated circuit. The variable resistor sets the voltage at which the LED turns on. Set the control so that the LED lights when the voltage from the car's ignition switch drops below 13.8 volts. The LED normally will light every now and then for a short period of time. But, if it stays on for very long, your electrical system is in trouble.

Fig. 10-21

OVERSPEED INDICATOR


14V
430Ω
68Ω
150Ω
8
7
6
5
CHARGE PUMP
+
+
f_{IN}
1
2
3
4
C_1
0.033μF
C_2
1.0μF
R_1
100K
30K
LED

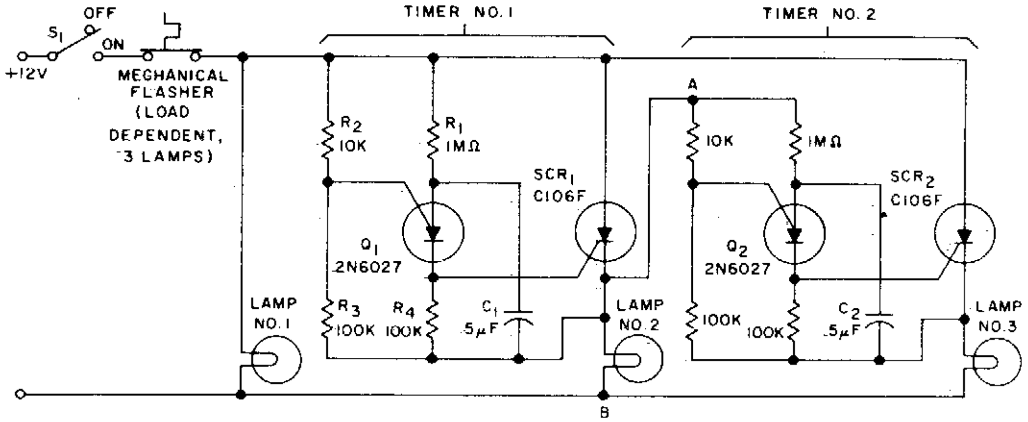
FLASHING BEGINS WHEN $f_{IN} > 100$ Hz
FLASH RATE INCREASES WITH INPUT FREQUENCY
INCREASE BEYOND TRIP POINT

Circuit Notes

An op-amp comparator is used to compare the converter output with a dc threshold voltage. The circuit flashes the LED when the input frequency exceeds 100 Hz. Increases in frequency raise the average current out of terminal 3 so that frequencies above 100 Hz reduce the charge time of C2, increasing the LED flashing rate. IC = LM2907 or LM2917

Fig. 10-22

SEQUENTIAL FLASHER FOR AUTOMOTIVE TURN SIGNALS


OFF
S_1
ON
+12V
MECHANICAL FLASHER (LOAD DEPENDENT, 3 LAMPS)
TIMER NO. 1
TIMER NO. 2
A
R_2 10K
R_1 1MΩ
10K
1MΩ
SCR_1 C106F
SCR_2 C106F
Q_1 2N6027
Q_2 2N6027
LAMP NO. 1
R_3 100K
R_4 100K
C_1 .5μF
LAMP NO. 2
100K
100K
C_2 .5μF
LAMP NO. 3
B

Circuit Notes

When the turn signal switch S1 is closed, lamp #1 will be activated and capacitor C1 will charge to the triggered voltage of Q1. As soon as the anode voltage on Q1 exceeds its gate voltage by 0.5 V, Q1 will switch into the low resistance mode, thereby triggering SCR1 to activate lamp #2 and the second timing circuit. After Q2 switches into the low resistance state, SCR2 will be triggered to activate lamp #3. When the thermal flasher interrupts the current to all three lamps, SCR1 and SCR2 are commutated and the circuit is ready for another cycle.

Fig. 10-23

AUTO LIGHTS-ON REMINDER

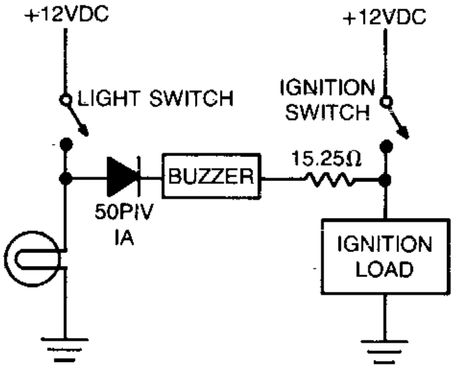

+12VDC
+12VDC
LIGHT SWITCH
IGNITION SWITCH
BUZZER
15.25Ω
50PIV
1A
IGNITION LOAD

Fig. 10-24

Circuit Notes

The alarm is composed of a diode, buzzer, and limiting resistor. The diode serves as a switch which allows the buzzer to sound off only when the light switch is closed and the ignition is turned off.

11

Battery Chargers

The sources of the following circuits are contained in the Sources section beginning on page 730. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

12 V Battery Charger
Simple Ni-Cad Battery Charger
12 V Battery Charger Control (20 Amps Rms Max.)
Battery Charger
Automatic Shutoff Battery Charger
200 mA-Hour, 12 V Ni-Cad Battery Charger
Ni-Cad Charger with Current and Voltage Limiting
Automotive Charger for Ni-Cad Battery Packs
Constant Voltage, Current-Limited Charger
Ni-Cad Charger
Simple Ni-Cad Battery Zapper
Battery Charging Regulator
Low-Cost Trickle Charger for 12V Storage Battery
Fast Charger for Ni-Cad Batteries
Current Limited 6 V Charger

12 V BATTERY CHARGER

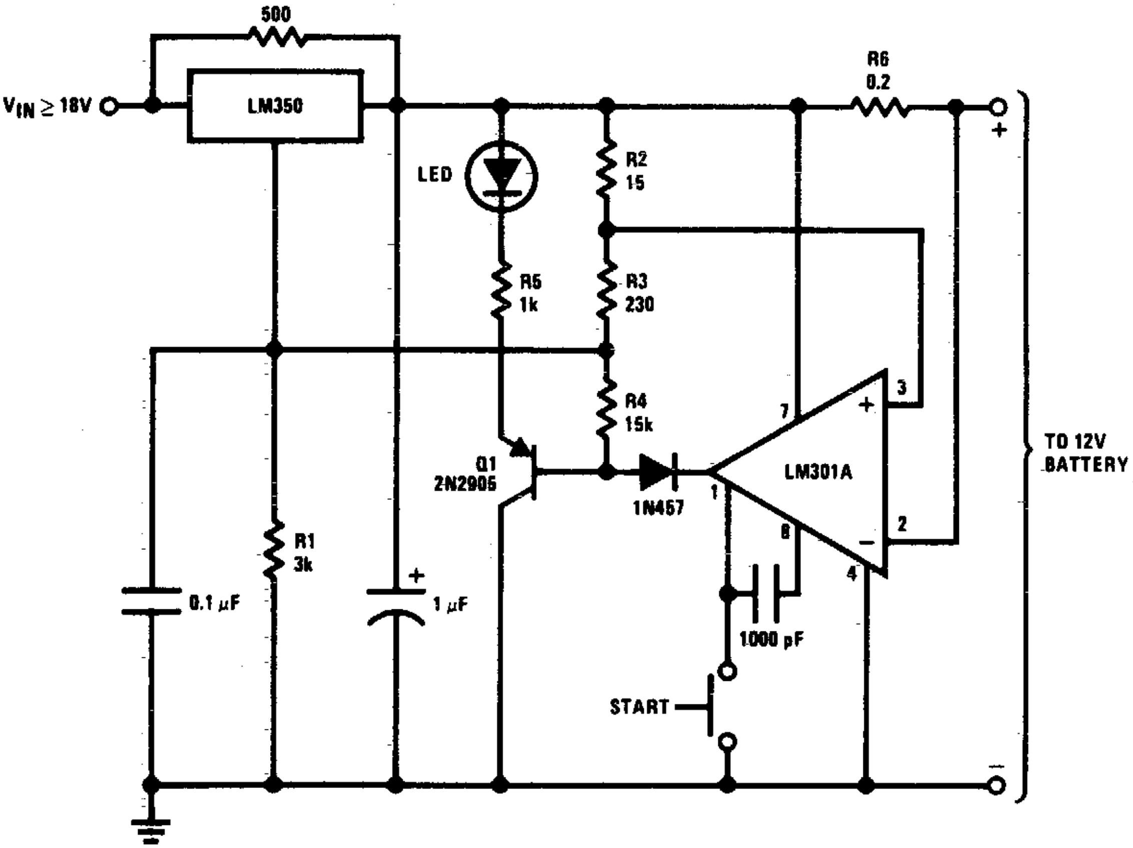

500
V_IN ≥ 18V
LM350
R6
0.2
+
LED
R2
15
R5
1k
R3
230
R4
15k
Q1
2N2905
1N457
LM301A
7
3
1
8
2
4
R1
3k
0.1 μF
1 μF
1000 pF
START
TO 12V
BATTERY
−

Fig. 11-1

Circuit Notes

This circuit is a high performance charger for gelled electrolyte lead-acid batteries. Charger quickly recharges battery and shuts off at full charge. Initially, charging current is limited to 2A. As the battery voltage rises, current to the battery decreases, and when the current has decreased to 150 mA, the charger switches to a lower float voltage preventing overcharge. When the start switch is pushed, the output of the charger goes to 14.5 V. As the battery approaches full charge, the charging current decreases and the output voltage is reduced from 14.5 V to about 12.5 V terminating the charging. Transistor Q1 then lights the LED as a visual indication of full charge.

SIMPLE NI-CAD BATTERY CHARGER

PARTS LIST FOR
NICAD BATTERY CHARGER

C1—100-μF, 50-V electrolytic capacitor
D1—1-A, 400 PIV-silicon rectifier
Q1—40-W, pnp power transistor
R1—2000-ohm potentiometer
T1—24-Vac, 117-Vac primary filament transformer

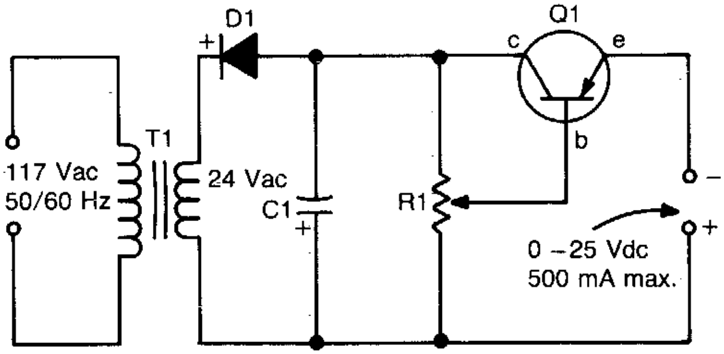
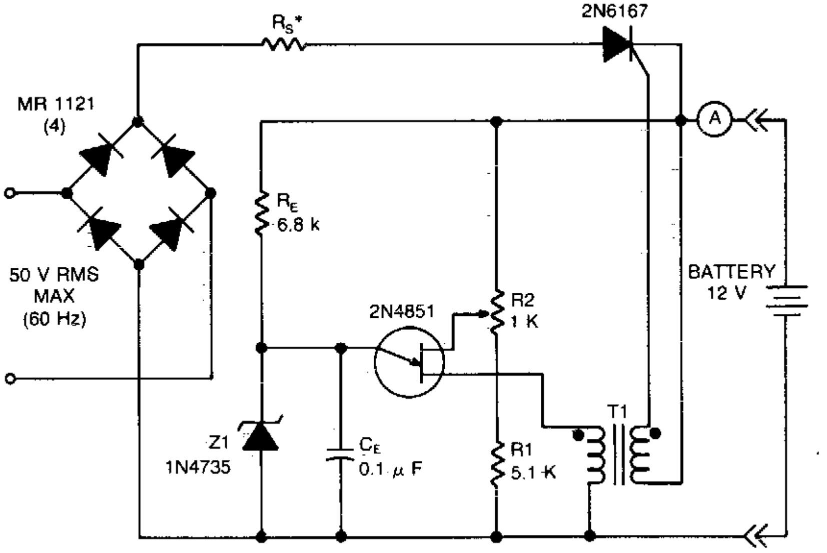

D1
+
Q1
c
e
b
T1
117 Vac
50/60 Hz
24 Vac
C1
+
R1
−
+
0 −25 Vdc
500 mA max.

Fig. 11-2

Circuit Notes

This circuit provides an adjustable output voltage up to 35 Vdc and maximum output current of 50 mA. Transistor Q1 dissipates quite a bit of heat and must be mounted on a heatsink.

12 V BATTERY CHARGER CONTROL (20 AMPS RMS MAX.)


R_S*
2N6167
MR 1121
(4)
A
R_E
6.8 k
50 V RMS
MAX
(60 Hz)
BATTERY
12 V
2N4851
R2
1 K
T1
Z1
1N4735
C_E
0.1 μ F
R1
5.1 K

T1 - PRIMARY = 30 TURNS #22
SECONDARY = 45 TURNS #22
CORE = FERROXCUBE 203 F 181-3C3
* R_S - SERIES RESISTANCE TO LIMIT CURRENT THROUGH SCR.
2N6167 IS RATED AT 20 AMPS RMS.

Fig. 11-3

BATTERY CHARGER

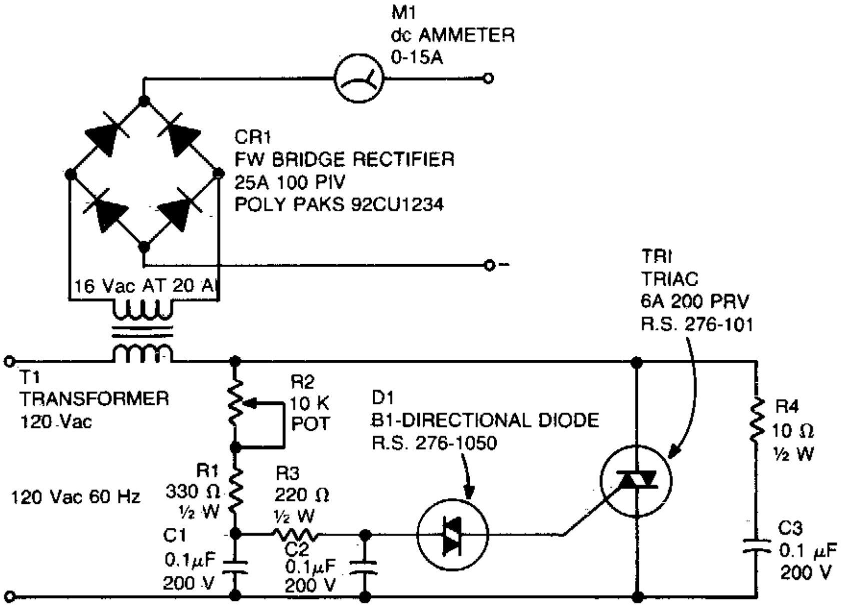

M1
dc AMMETER
0-15A
CR1
FW BRIDGE RECTIFIER
25A 100 PIV
POLY PAKS 92CU1234
16 Vac AT 20 A
TRI
TRIAC
6A 200 PRV
R.S. 276-101
T1
TRANSFORMER
120 Vac
R2
10 K
POT
D1
BI-DIRECTIONAL DIODE
R.S. 276-1050
R4
10 Ω
½ W
R1
330 Ω
½ W
R3
220 Ω
½ W
120 Vac 60 Hz
C1
0.1μF
200 V
C2
0.1μF
200 V
C3
0.1 μF
200 V
-

Fig. 11-4

Circuit Notes

A diac is used in the gate circuit to provide a threshold level for firing the triac. C3 and R4 provide a transient suppression network. R1, R2, R3, C1, and C2 provide a phase-shift network for the signal being applied to the gate. R1 is selected to limit the maximum charging current at full rotation of R2.

AUTOMATIC SHUTOFF BATTERY CHARGER

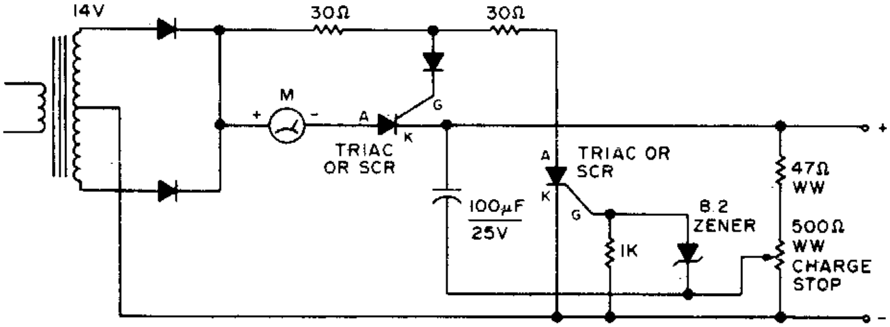

14V
30Ω
30Ω
M
+
-
A
G
K
TRIAC
OR SCR
A
TRIAC OR
SCR
K
G
100μF
25V
8.2
ZENER
1K
47Ω
WW
500Ω
WW
CHARGE
STOP
+
-

Fig. 11-5

Circuit Notes

Adjust by setting the 500 ohm resistor while attached to a fully charged battery.

200 mA-HOUR, 12 V NI-CAD BATTERY CHARGER

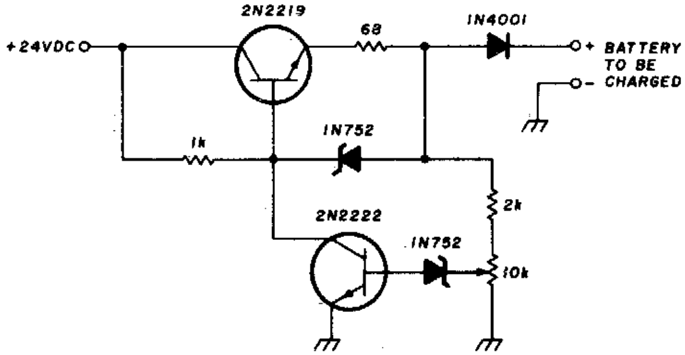

2N2219
68
1N4001
+24VDC
+ BATTERY TO BE CHARGED
-
1N752
1k
2k
2N2222
1N752
10k

Fig. 11-6

Circuit Notes

This circuit charges the battery at 75 mA until the battery is charged, then it reduces the current to a trickle rate. It will completely recharge a dead battery in four hours and the battery can be left in the charger indefinitely. To set the shut-off point, connect a 270-ohm, 2-watt resistor across the charge terminals and adjust the pot for 15.5 volts across the resistor.

NI-CAD CHARGER WITH CURRENT AND VOLTAGE LIMITING

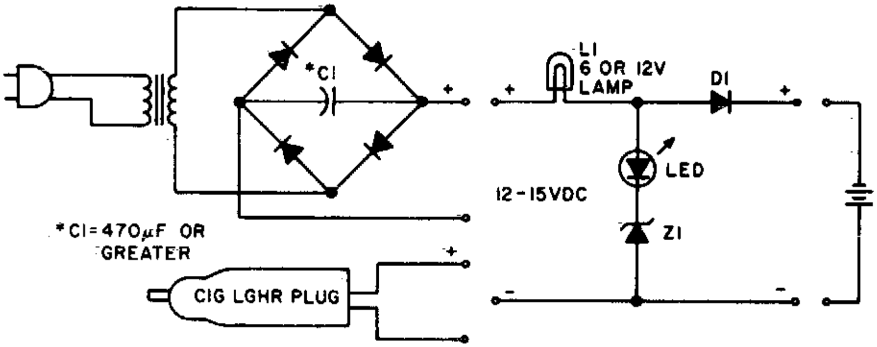

*C1
+
+
L1
6 OR 12V LAMP
D1
+
LED
12-15VDC
Z1
*C1=470μF OR GREATER
+
CIG LGHR PLUG
-
-

Fig. 11-7

Circuit Notes

Lamp L1 will glow brightly and the LED will be out when the battery is low and being charged, but the LED will be bright and the light bulb dim when the battery is almost ready. L1 should be a light bulb rated for the current you want (usually the battery capacity divided by 10). Diode D1 should be at least 1 A, and Z1 is a 1 W zener diode with a voltage determined by the full-charge battery voltage minus 1.5 V. After the battery is fully charged, the circuit will float it at about battery capacity divided by 100 mA.

AUTOMOTIVE CHARGER FOR NI-CAD BATTERY PACKS

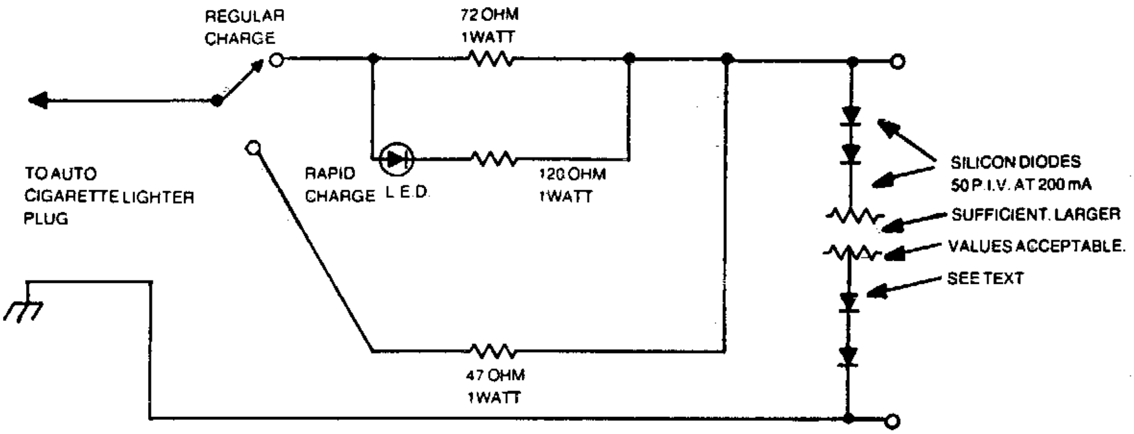

REGULAR
CHARGE
72 OHM
1WATT
TO AUTO
CIGARETTE LIGHTER
PLUG
RAPID
CHARGE
L.E.D.
120 OHM
1WATT
SILICON DIODES
50 P.I.V. AT 200 mA
SUFFICIENT. LARGER
VALUES ACCEPTABLE.
SEE TEXT
47 OHM
1WATT

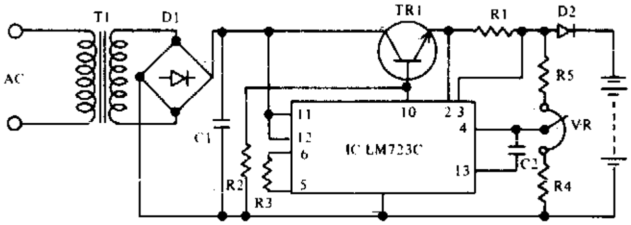
Fig. 11-8

Circuit Notes

The number of silicon diodes across the output is determined by the voltage of the battery pack. Figure each diode at 0.7 volt. For example, a 10.9- volt pack would require 10.9/0.7 = 15.57, or 16 diodes.

CONSTANT-VOLTAGE, CURRENT-LIMITED CHARGER

IC LM723C VOLTAGE REGULATOR (FOR 12V dc OUTPUT 0.42A MAX.)


T1
D1
AC
TR1
R1
D2
R5
VR
C1
C2
R2
R3
R4
IC LM723C
11
12
6
5
10
2 3
4
13

T1	TRANSFORMER, DC 13V (RMS), 1-3A (RMS)
D1, D2	100V 1A DIODE
C1	50V, 470μF ELECTROLYTIC CONDENSER
TR1	MJ2840 10A 60V 150W (MOTOROLA)
IC	LM723C (NATIONAL SEMICONDUCTOR)
R1	4.7 OHM 1/2W 3P
R2	5.1K OHM 1/4W
R3	3.9K OHM 1/4W
R4	7.5K OHM 1/4W
R5	8.2K OHM 1/4W
VR	2K OHM
C2	50V 1000PF

Circuit Notes

For 12 V sealed lead-acid batteries.

Fig. 11-9

NI-CAD CHARGER

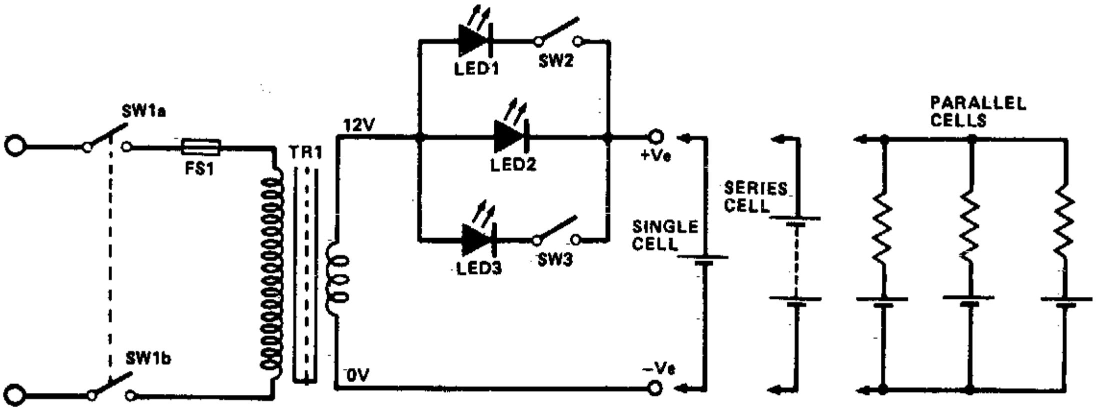

LED1
SW2
SW1a
12V
TR1
FS1
LED2
+Ve
SERIES CELL
PARALLEL CELLS
SINGLE CELL
LED3
SW3
SW1b
0V
−Ve

Fig. 11-10

Circuit Notes

This circuit uses constant current LEDs to adjust charging current. It makes use of LEDs that pass a constant current of about 15 mA for an applied voltage range of 2-18 V. They can be paralleled to give any multiple of 15 mA and they light up when current is flowing. The circuit will charge a single cell at 15, 30 or 45 mA or cells in series up to the rated supply voltage limit (about 14 V).

SIMPLE NI-CAD BATTERY ZAPPER

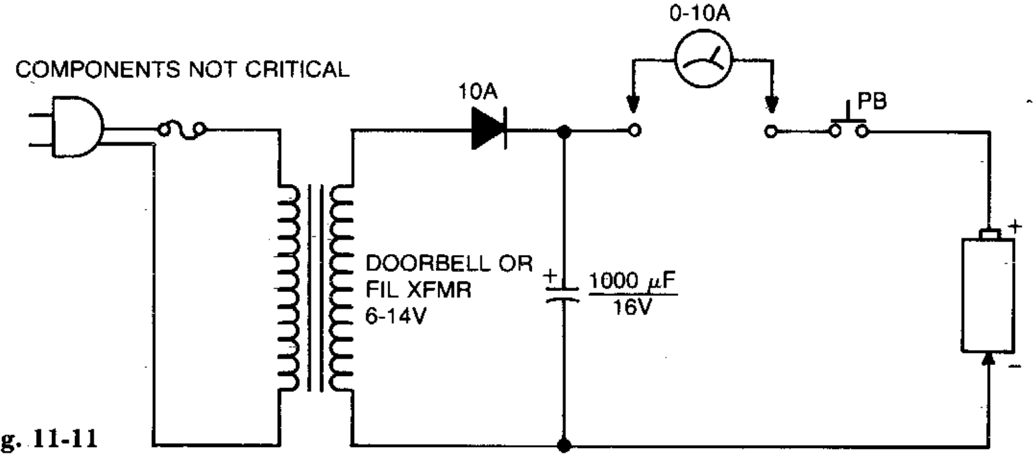
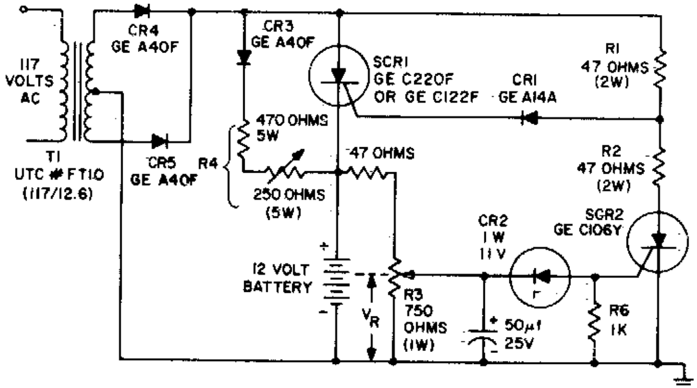

0-10A
COMPONENTS NOT CRITICAL
10A
PB
+
DOORBELL OR FIL XFMR 6-14V
+
1000 μF 16V
−

Fig. 11-11

Circuit Notes

This circuit is used to clear internal shorts in nickel cadmium batteries. To operate, connect ni-cad to output and press the pushbutton for three seconds.

BATTERY CHARGING REGULATOR


117 VOLTS AC
T1
UTC # FT10
(117/12.6)
CR4
GE A40F
CR5
GE A40F
CR3
GE A40F
R4
470 OHMS
5W
250 OHMS
(5W)
SCR1
GE C220F
OR GE C122F
CR1
GE A14A
R1
47 OHMS
(2W)
47 OHMS
R2
47 OHMS
(2W)
SCR2
GE C106Y
CR2
1 W
11 V
+
12 VOLT
BATTERY
-
V_R
R3
750
OHMS
(1W)
+ 50μf
25V
R6
1K

ALL RESISTORS 1/2 WATT EXCEPT AS NOTED

Fig. 11-12

Circuit Notes

The circuit is capable of charging a 12-volt battery at up to a six ampere rate. Other voltages and currents, from 6 to 600 volts and up to 300 amperes, can be accommodated by suitable component selection. When the battery voltage reaches its fully charged level, the charging SCR shuts off, and a trickle charge as determined by the value of R4 continues to flow.

LOW-COST TRICKLE CHARGER FOR 12 V STORAGE BATTERY

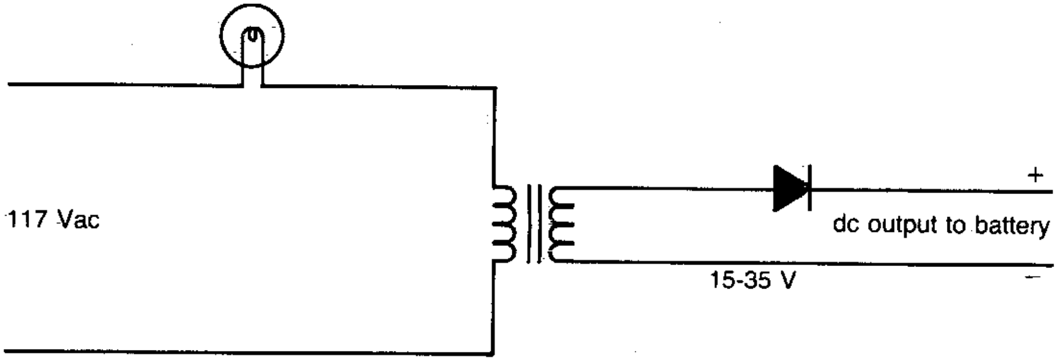
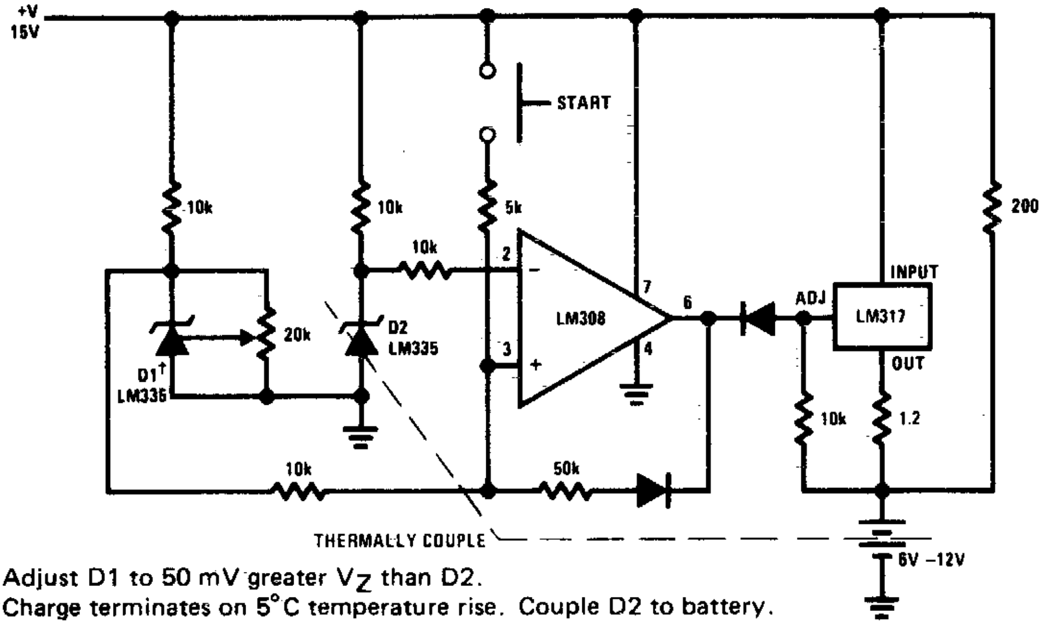

117 Vac
+
dc output to battery
15-35 V
−

Fig. 11-13

Circuit Notes

Charge rate can be varied and is based on the size of bulb.

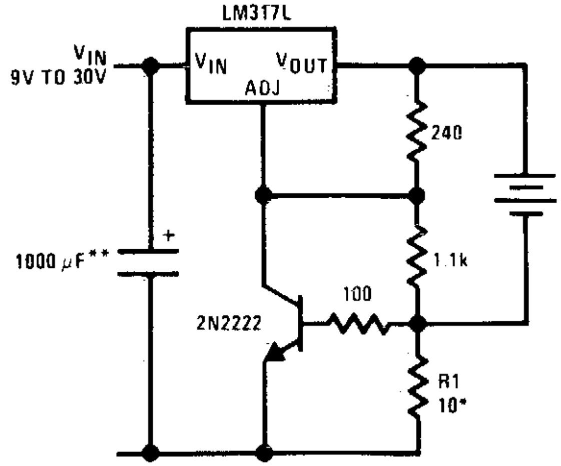
FAST CHARGER FOR NI-CAD BATTERIES


+V
15V
START
10k
10k
5k
200
10k
2
3
7
6
4
LM308
ADJ
INPUT
LM317
OUT
20k
D1†
LM336
D2
LM335
10k
1.2
10k
50k
THERMALLY COUPLE
6V -12V

Adjust D1 to 50 mV greater V_Z than D2.
Charge terminates on 5°C temperature rise. Couple D2 to battery.

Fig. 11-14

CURRENT LIMITED 6 V CHARGER


LM317L
V_{IN}
9V TO 30V
V_{IN}
V_{OUT}
ADJ
240
1000 μF**
+
1.1k
100
2N2222
R1
10*

* Sets peak current, $I_{PEAK} = 0.6V/R1$
* * 1000 μF is recommended to filter
out any input transients.

Fig. 11-15

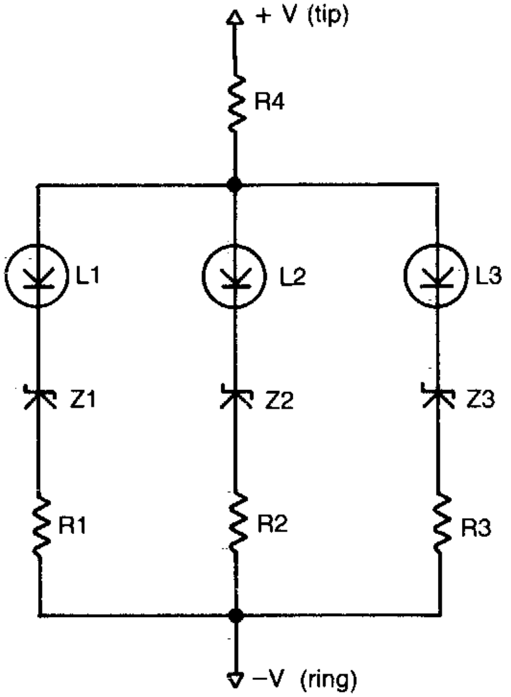
12

Battery Monitors

The sources of the following circuits are contained in the Sources section beginning on page 730. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

Solid-State Battery Voltage Indicator
Ni-Cad Discharge Limiter
Battery Condition Indicator
Equipment on Reminder
Battery Charge/Discharge Indicator
Precision Battery Voltage Monitor for HTs
Low Voltage Monitor
Undervoltage indicator for Battery Operated Equipment
Low Battery Indicator
Battery-Level Indicator
Battery-Threshold Indicator

SOLID-STATE BATTERY VOLTAGE INDICATOR

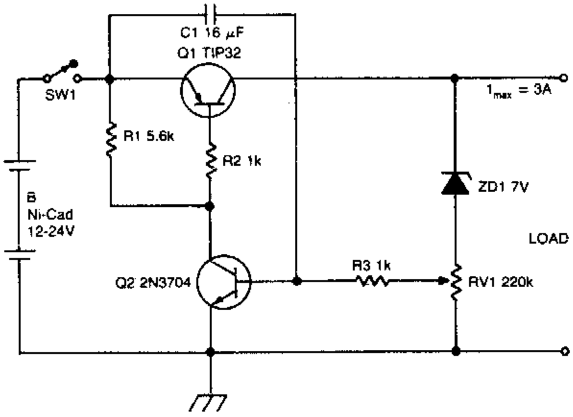

+ V (tip)
R4
L1
L2
L3
Z1
Z2
Z3
R1
R2
R3
−V (ring)

R1, R2, R3 = 47 Ω
R4 = 39 Ω
Z1 = 9.8 volt zener diode
Z2 = 11.1 volt zener diode
Z3 = 11.5 volt zener diode
L1 – L3 = light emitting diodes

Two lights on - OK (L1 + L2)
One light on - low voltage (L1 only)
Three lights on - overvoltage (L1 + L2 + L3)

Fig. 12-1

NI-CAD DISCHARGE LIMITER


C1 16 µF
Q1 TIP32
SW1
I_{max} = 3A
R1 5.6k
R2 1k
ZD1 7V
B
Ni-Cad
12-24V
LOAD
R3 1k
Q2 2N3704
RV1 220k

Circuit Notes

The circuit disconnects the battery from the load when output voltage falls below a preset level. C1 charges through R1 and turns on Q2. Collector current flows through R2 turning Q1 on and battery is connected to the load. When the output voltage falls below a point set by RV1, Q2 turns off, Q1 turns off and further discharge of the battery is prevented.

Fig. 12-2

BATTERY CONDITION INDICATOR

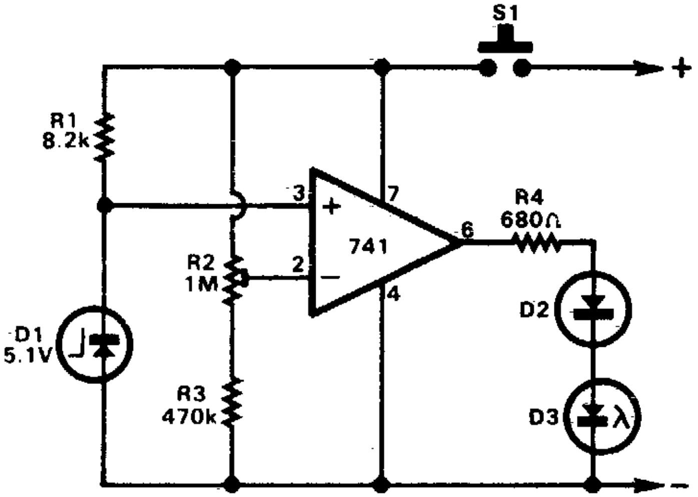

S1
+
R1
8.2k
3
+
7
741
6
R4
680Ω
R2
1M
2
−
4
D1
5.1V
D2
R3
470k
D3
−

Fig. 12-3

Circuit Notes

A 741 op amp is employed as a voltage comparator. The noninverting input is connected to zener reference source. Reference voltage is 5.1V. R2 is adjusted so that the voltage at the inverting input is half the supply voltage. When supply is higher than 10.2V, the LED will not light. When the supply falls just fractionally below the 10.2V level, the IC inverting input will be slightly negative of the noninverting input, and the output will swing fully positive. The LED will light, indicating that the supply voltage has fallen to the preset threshold level. The LED can be made to light at other voltages by adjusting R2.

EQUIPMENT ON REMINDER

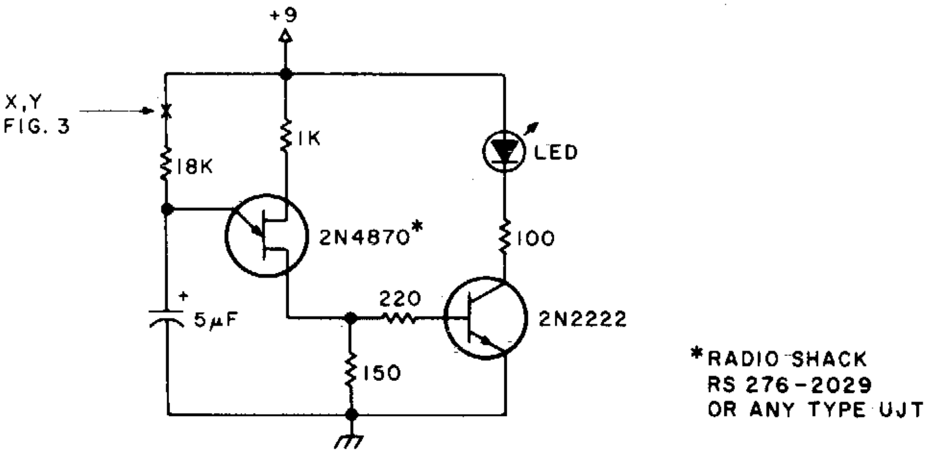

+9
X,Y
FIG. 3
18K
1K
LED
2N4870*
100
+
5μF
220
2N2222
150
*RADIO SHACK
RS 276-2029
OR ANY TYPE UJT

Fig. 12-4

Circuit Notes

Due to the low duty cycle of flashing LED, the average current drain is 1 mA or less.

BATTERY CHARGE/DISCHARGE INDICATOR

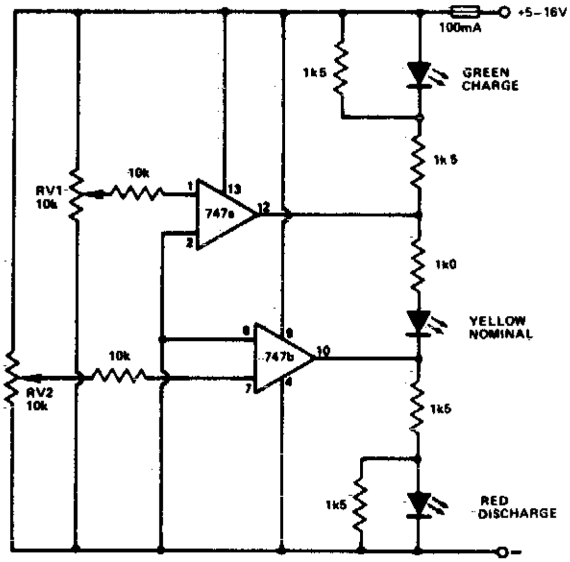

+5–16V
100mA
1k5
GREEN CHARGE
1k5
RV1 10k
10k
1
13
747a
12
2
1k0
YELLOW NOMINAL
8
9
747b
10
10k
7
4
RV2 10k
1k5
1k5
RED DISCHARGE

Fig. 12-5

Circuit Notes

This circuit monitors car battery voltage. It provides an indication of nominal supply voltage as well as low or high voltage. RV1 and RV2 adjust the point at which the red/yellow and yellow/green LEDs are on or off. For example the red LED comes on at 11V, and the green LED at 12V. The yellow LED is on between these values.

PRECISION BATTERY VOLTAGE MONITOR FOR HTS

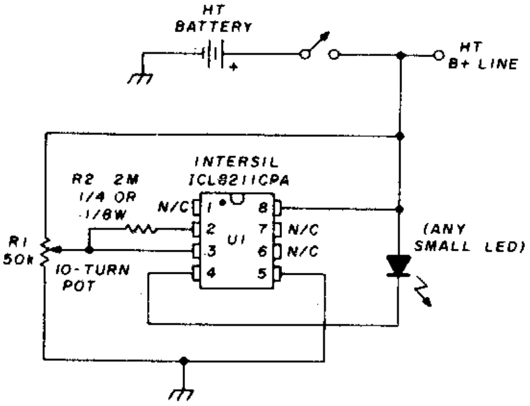

HT BATTERY
+
HT B+ LINE
INTERSIL ICL8211CPA
R2 2M 1/4 OR 1/8W
N/C
1
2
3
4
5
6
7
8
U1
N/C
N/C
R1 50k
10-TURN POT
(ANY SMALL LED)

Fig. 12-6

Circuit Notes

The precision voltage-monitor chip contains a temperature-compensated voltage reference. R1 divides down the battery voltage to match the built-in reference voltage of IC1 (1.15 volts). When the voltage at pin 3 falls below 1.15 volts, pin 4 supplies a constant current of 7 mA to drive a small LED. About 0.2 volt of hysteresis is added with R2. Without hysteresis, the LED could flicker on and off when the monitored voltage varies around the set point, as might be the case on voice peaks during receive.

LOW-VOLTAGE MONITOR

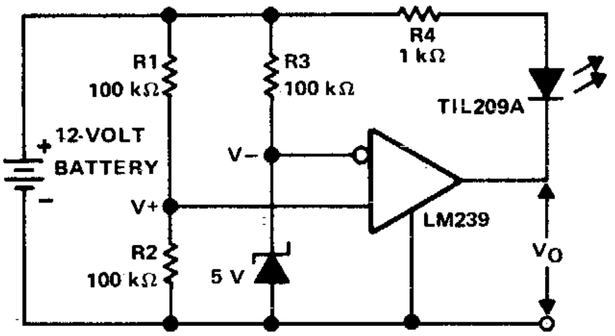

R1
100 kΩ
R3
100 kΩ
R4
1 kΩ
TIL209A
12-VOLT BATTERY
+
-
V-
V+
LM239
R2
100 kΩ
5 V
V_O

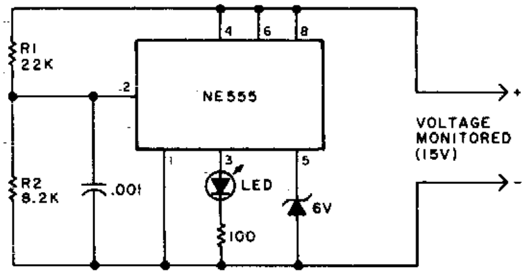
Fig. 12-7

a. SCHEMATIC OF CIRCUIT FOR LOW-VOLTAGE INDICATOR

Circuit Notes

This circuit monitors the voltage of a battery and warns the operator when the battery voltage is below a preset level by turning on an LED. The values are set for a 12V automobile battery. The preset value is 10 volts.

UNDERVOLTAGE INDICATOR FOR BATTERY OPERATED EQUIPMENT


R1
22K
R2
8.2K
.001
NE555
4
6
8
2
1
3
5
LED
6V
100
VOLTAGE MONITORED (15V)
+
-

Circuit Notes

Due to the low duty cycle of flashing LED, the average current drain is 1 mA or less. The NE555 will trigger the LED on when the monitored voltage falls to 12 volts. The ratio of R1 to R2 only needs to be changed if it is desired to change the voltage point at which the LED is triggered.

Fig. 12-8

LOW BATTERY INDICATOR

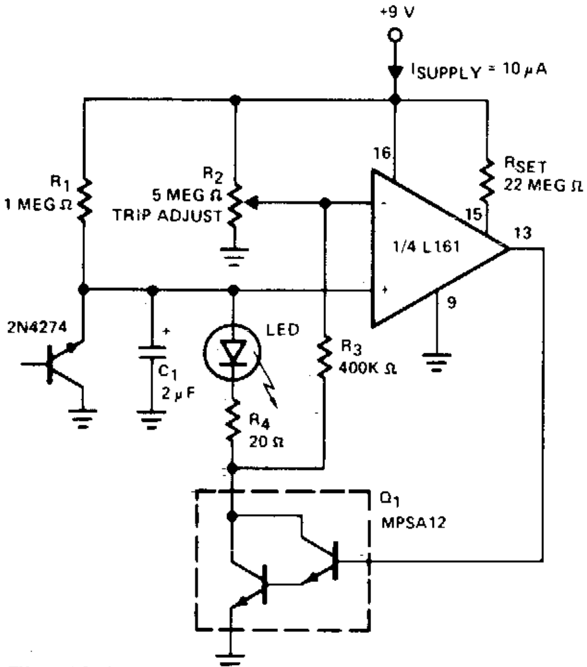

+9 V
I_SUPPLY = 10 µA
16
R_SET
22 MEG Ω
R1
1 MEG Ω
R2
5 MEG Ω
TRIP ADJUST
15
13
1/4 L161
9
2N4274
+
C1
2 µF
LED
R3
400K Ω
R4
20 Ω
Q1
MPSA12

Fig. 12-9

Circuit Notes

The indicator flashes an LED when the battery voltage drops below a certain threshold. 2N4274 emitter-base junction serves as a zener which establishes about 6V on the L161's positive input. As the battery drops, the L161 output goes high. This turns on the Darlington, which discharges C1 through the LED. The interval between flashes is roughly two seconds and gives a low battery warning with only 10 μA average power drain.

BATTERY-LEVEL INDICATOR

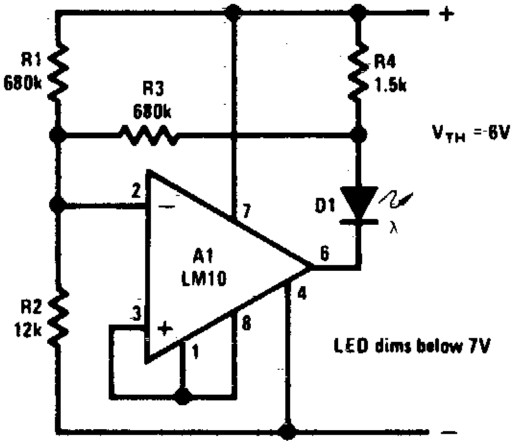

+
R1
680k
R3
680k
R4
1.5k
V_TH = 6V
2
7
D1
λ
A1
LM10
6
4
R2
12k
3
8
1
LED dims below 7V
−

Fig. 12-10

BATTERY-THRESHOLD INDICATOR

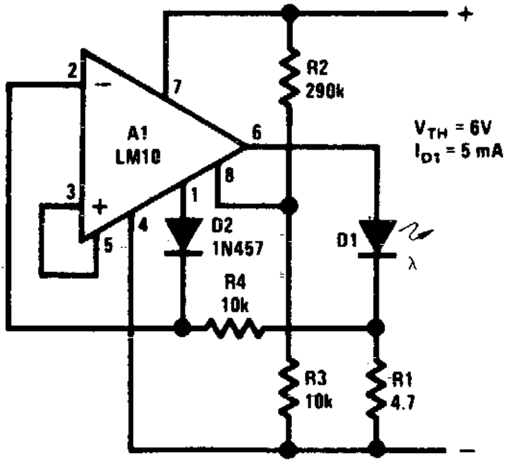

2
7
+
R2
290k
A1
LM10
6
V_TH = 6V
I_D1 = 5 mA
8
3
1
4
5
D2
1N457
D1
λ
R4
10k
R3
10k
R1
4.7
−

Fig. 12-11

13

Buffers

The sources of the following circuits are contained in the Sources section beginning on page 730. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

Sine Wave Output Buffer Amplifier
Single-Supply AC Buffer Amplifier
Single-Supply AC Buffer
High-Speed 6-Bit A/D Buffer
High Impedance, Low Capacitance
Wideband Buffer
High Resolution ADC Input Buffer
100 × Buffer Amplifier
10 × Buffer Amplifier
Stable High Impedance Buffer
High-Speed Single Supply AC Buffer

SINE WAVE OUTPUT BUFFER AMPLIFIER

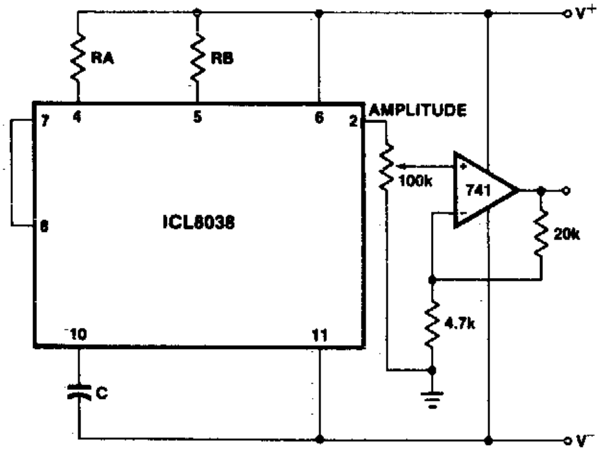

V+
RA
RB
7
4
5
6
2
AMPLITUDE
100k
741
20k
8
ICL8038
4.7k
10
11
C
V−

Fig. 13-1

Circuit Notes

The sine wave output has a relatively high output impedance (1K typ). The circuit provides buffering, gain, and amplitude adjustment. A simple op amp follower could also be used.

SINGLE SUPPLY AC BUFFER AMPLIFIER

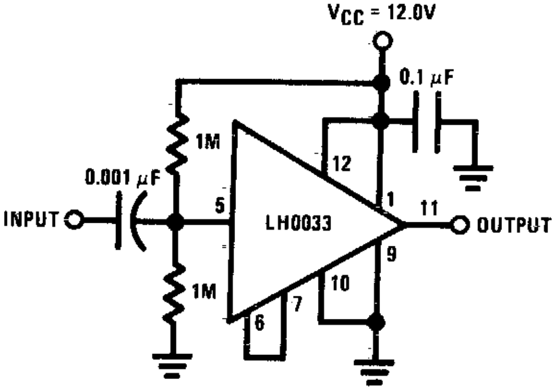
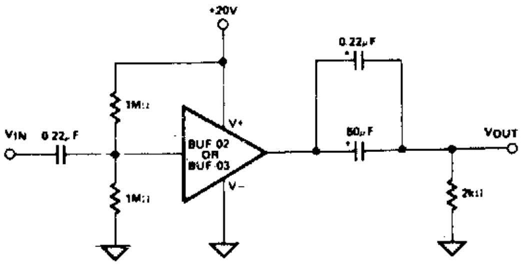

V_{CC} = 12.0V
0.1 μF
1M
0.001 μF
12
INPUT
5
LH0033
1
11
OUTPUT
9
1M
10
7
6

Fig. 13-2

Circuit Notes

The input is dc biased to mid-operating point and is ac coupled. Its input impedance is approximately 500K at low frequencies. For dc loads referenced to ground, the quiescent current is increased by the load current set at the input dc bias voltage.

SINGLE SUPPLY AC BUFFER (HIGH SPEED)


+20V
0.22µF
*
1MΩ
V+
VIN
0.22µF
BUF-02
OR
BUF-03
60µF
*
VOUT
1MΩ
V−
2kΩ

*NEEDED FOR LOW IMPEDANCE AT HIGH FREQUENCIES

f_{LOW} AT V_{IN} = 1.46Hz }
f_{LOW} AT V_{OUT} = 1.59Hz } −3dB

ASSUME V_{IN} = 10V P-P SINE WAVE (5V-PEAK)
THEN FULL POWER BANDWIDTH IS
796kHz FOR BUF-02, AND
9.55MHz FOR BUF-03

Fig. 13-3

HIGH SPEED 6-BIT A/D BUFFER

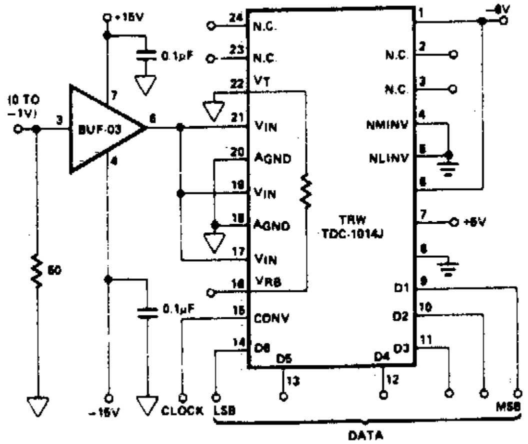

+15V
0.1µF
(0 TO −1V)
3
7
BUF-03
6
4
60
0.1µF
−15V
CLOCK
LSB
24
N.C.
23
N.C.
22
VT
21
VIN
20
AGND
19
VIN
18
AGND
17
VIN
16
VRB
15
CONV
14
D6
D5
13
TRW
TDC-1014J
D4
12
1
−6V
2
N.C.
3
N.C.
4
NMINV
5
NLINV
6
7
+6V
8
9
D1
10
D2
11
D3
MSB
DATA

Fig. 13-4

HIGH IMPEDANCE LOW CAPACITANCE WIDEBAND BUFFER

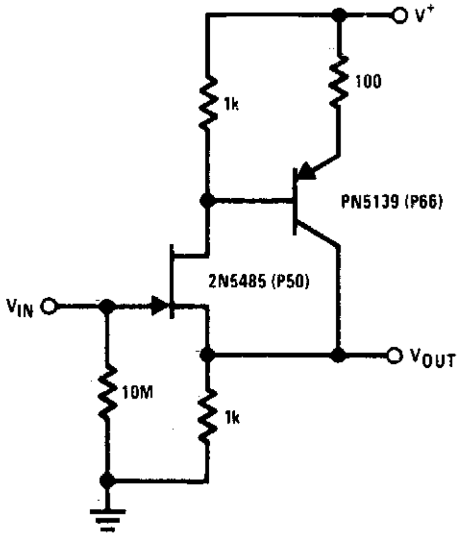
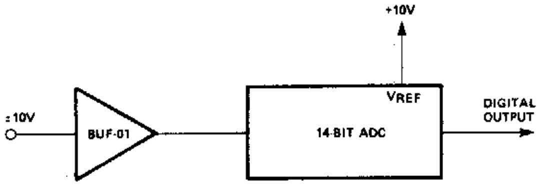

V+
100
1k
PN5139 (P66)
2N5485 (P50)
VIN
VOUT
10M
1k

Fig. 13-5

Circuit Notes

The 2N5485 has low input capacitance which makes this compound series-feedback buffer a wide-band unity gain amplifier.

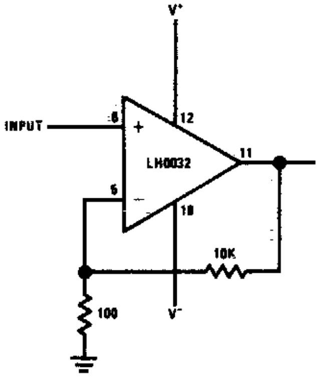
HIGH RESOLUTION ADC INPUT BUFFER


+10V
VREF
±10V
BUF-01
14-BIT ADC
DIGITAL
OUTPUT

- MAXIMUM ERROR FROM BUF-01 IS 300µV.
- RESOLUTION OF 10V, 14-BIT ADC IS 610µV.
- BUF-01 RESOLVES 1/2 LSB OF 14-BIT SYSTEM.

Fig. 13-6

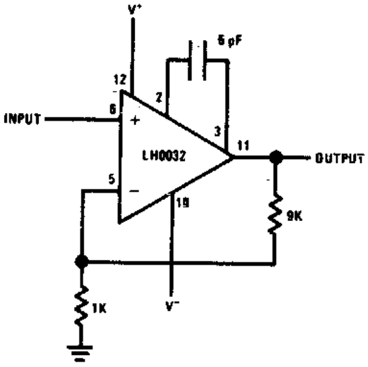
100 × BUFFER AMPLIFIER


V+
INPUT
6
+
12
LH0032
11
5
−
10
10K
100
V−

TYP. BW3dB = 5 MHz

Fig. 13-7

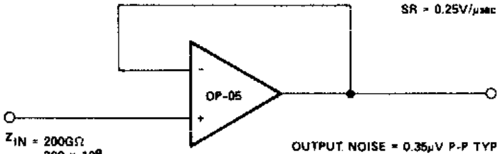
10 × BUFFER AMPLIFIER


V+
5 pF
12
2
6
INPUT
+
3
LH0032
11
OUTPUT
5
−
10
9K
1K
V−

TYP. BW3dB = 10 MHz

Fig. 13-8

STABLE, HIGH IMPEDANCE BUFFER



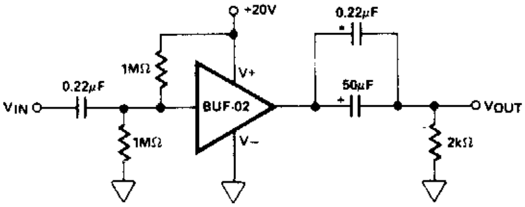
SR = 0.25V/μsec

Z_{IN} = 200GΩ
200 × 10^9
I_N = ±1.0nA

OUTPUT NOISE = 0.35μV P-P TYP
OUTPUT OFFSET = 0.2mV TYP
BANDWIDTH = 1.2MHz

Fig. 13-9

HIGH-SPEED SINGLE-SUPPLY AC BUFFER


+20V
0.22μF
*
1MΩ
V+
0.22μF
50μF
+
VIN
BUF-02
VOUT
1MΩ
V−
2kΩ

*NEEDED FOR LOW IMPEDANCE AT HIGH FREQUENCIES

f_{LOW} AT V_{IN} = 1.45Hz
f_{LOW} AT V_{OUT} = 1.59Hz −3dB
ASSUME V_{IN} = 10V P-P SINE WAVE (5V PEAK)
THEN FULL POWER BANDWIDTH IS
APPROXIMATELY 800kHz.

Fig. 13-10

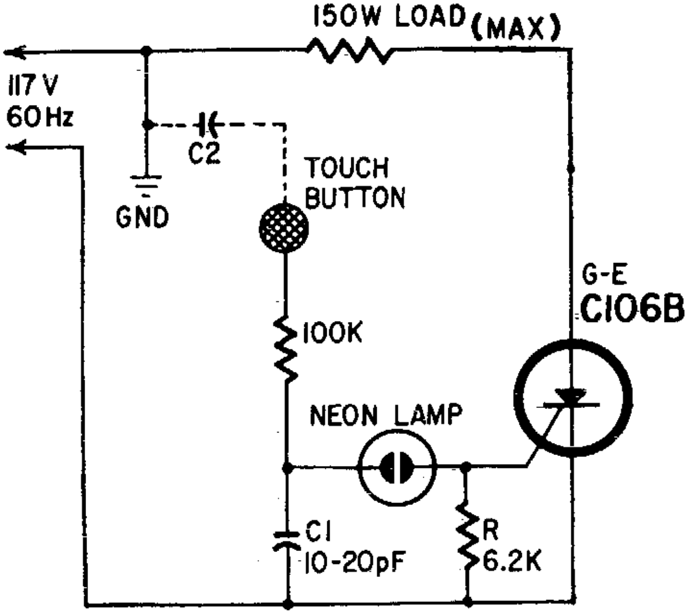
14

Capacitance (Touch) Operated Circuits

The sources of the following circuits are contained in the Sources section beginning on page 730. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

- Capacitance Relay
- Capacitance Operated, Battery Powered Light
- Touch Sensitive Switch
- Low Current Touch Switch
- Capacitance Switched Light
- Momentary Operation Touch Switch
- Touch Triggered Bistable
- Capacitance Operated Alarm to Foil Purse Snatchers
- Self-Biased Proximity Sensor Works on Detected Changing Fields
- Touch Switch or Proximity Detector
- Finger Touch Touch or Control Switch
- Proximity Detector
- Touch Circuit
- CMOS Touch Switch
- Latching Double-Button Touch Switch

CAPACITANCE RELAY



NOTE: ALL RESISTORS 1/2 WATT

Fig. 14-1

Circuit Notes

Capacitor C1 and body capacitance (C2) of the operator form the voltage divider from the hot side of the ac line to ground. The voltage across C1 is determined by the ratio of C1 to C2. The higher voltage is developed across the smaller capacitor. When no one is close to the touch button, C2 is smaller than C1. When a hand is brought close to the button, C2 is many times larger than C1 and the major portion of the line voltage appears across C1. This voltage fires the neon lamp, C1 and C2 discharge through the SCR gate, causing it to trigger and pass current through the load. The sensitivity of the circuit depends on the area of the touch plate. When the area is large enough, the circuit responds to the proximity of an object rather than to touch. C1 may be made variable so sensitivity can be adjusted.

CAPACITANCE OPERATED, BATTERY POWERED LIGHT

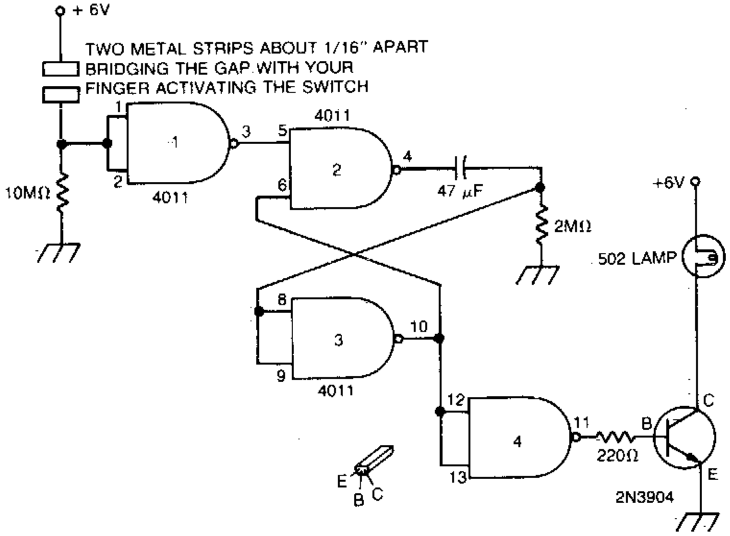
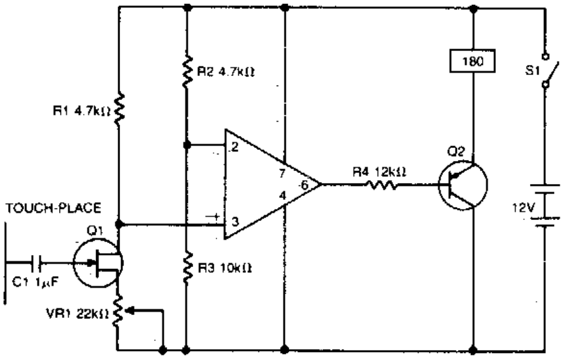

+ 6V
TWO METAL STRIPS ABOUT 1/16" APART BRIDGING THE GAP WITH YOUR FINGER ACTIVATING THE SWITCH
1
2
1
3
4011
5
6
2
4
4011
47 μF
10MΩ
2MΩ
+6V
502 LAMP
8
9
3
10
4011
12
13
4
11
220Ω
B
C
E
2N3904
E
B
C

Fig. 14-2

Circuit Notes

Touch the plate and the light will go on and remain on for a time determined by the time constant of the 47 μF capacitor and the 2M resistor.

TOUCH-SENSITIVE SWITCH

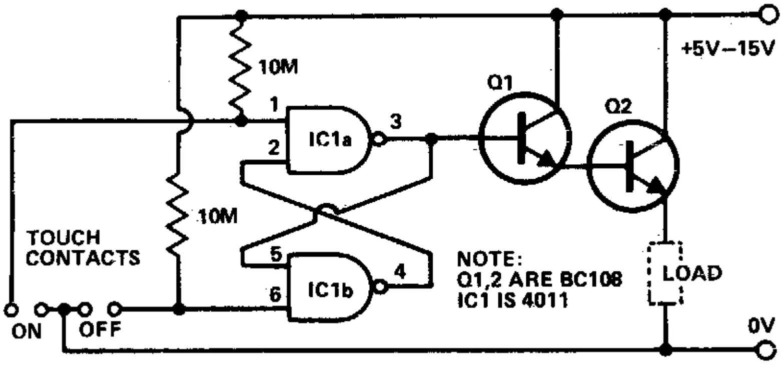

R2 4.7kΩ
180
S1
R1 4.7kΩ
2
7
6
4
3
+
R4 12kΩ
Q2
TOUCH-PLACE
Q1
12V
C1 1μF
R3 10kΩ
VR1 22kΩ

Circuit Notes

A high impedance input is provided by Q1, a general purpose field effect transistor. 741 op amp is used as a sensitive voltage level switch which in turn operates the current Q2, a medium current PNP bipolar transistor, thereby energizing the relay which can be used to control equipment, alarms, etc.

Fig. 14-3

LOW CURRENT TOUCH SWITCH



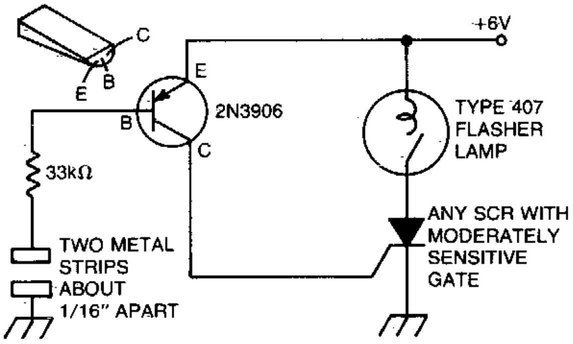
10M
1
2
3
IC1a
Q1
Q2
+5V–15V
10M
TOUCH CONTACTS
5
6
4
IC1b
NOTE:
Q1,2 ARE BC108
IC1 IS 4011
LOAD
ON
OFF
0V

Fig. 14-4

Circuit Notes

Touching the on contacts with a finger brings pin 3 high, turning on the Darlington pair and supplying power to the load (transistor radio etc). Q1 must be a high gain transistor, and Q2 is chosen for the current required by the load circuit.

CAPACITANCE SWITCHED LIGHT



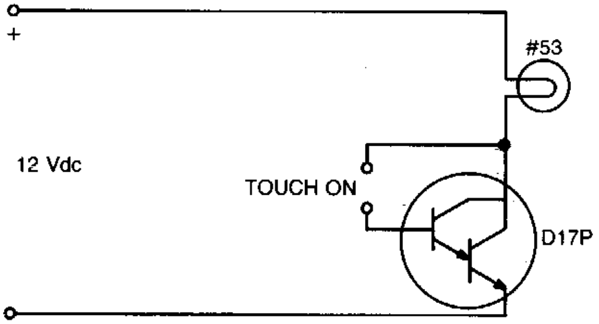
C
E
B
E
B
C
2N3906
+6V
TYPE 407 FLASHER LAMP
33kΩ
ANY SCR WITH MODERATELY SENSITIVE GATE
TWO METAL STRIPS ABOUT 1/16" APART

Fig. 14-5

Circuit Notes

The battery-powered light turns on easily, stays on for just a few seconds, and then turns off again. The circuit is triggered when you place a finger across the gap between two strips of metal, about 1/16th inch apart. Enough current will flow through your finger to trigger the SCR after being amplified by the 2N3906. Once the SCR is fired, current will flow through the bulb until its internal bimetal switch turns it off. Once that happens, the SCR will return to its nonconducting state.

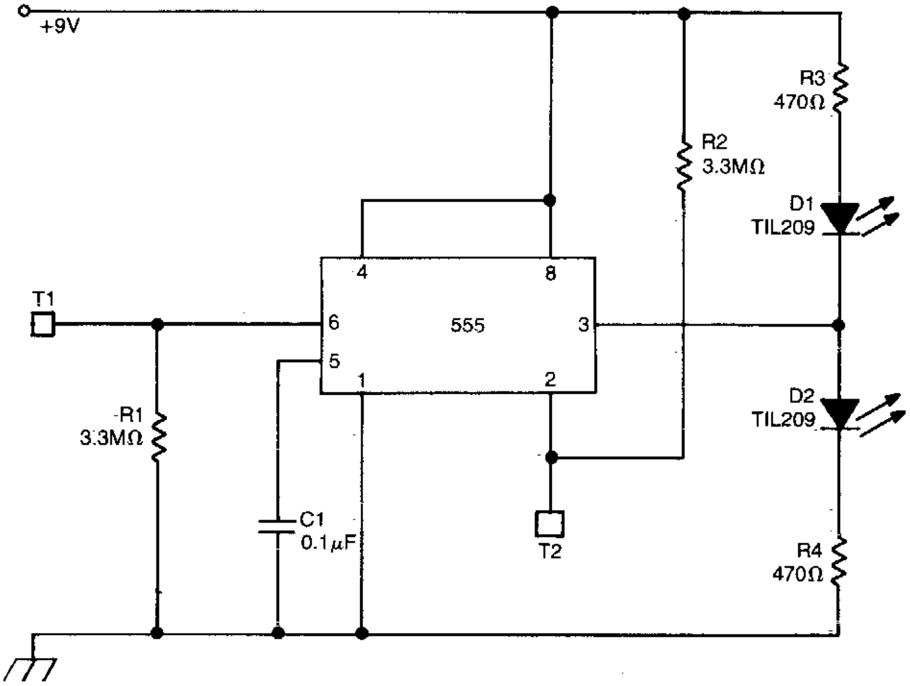
MOMENTARY OPERATION TOUCH SWITCH



+
#53
12 Vdc
TOUCH ON
D17P

Fig. 14-6

TOUCH TRIGGERED BISTABLE



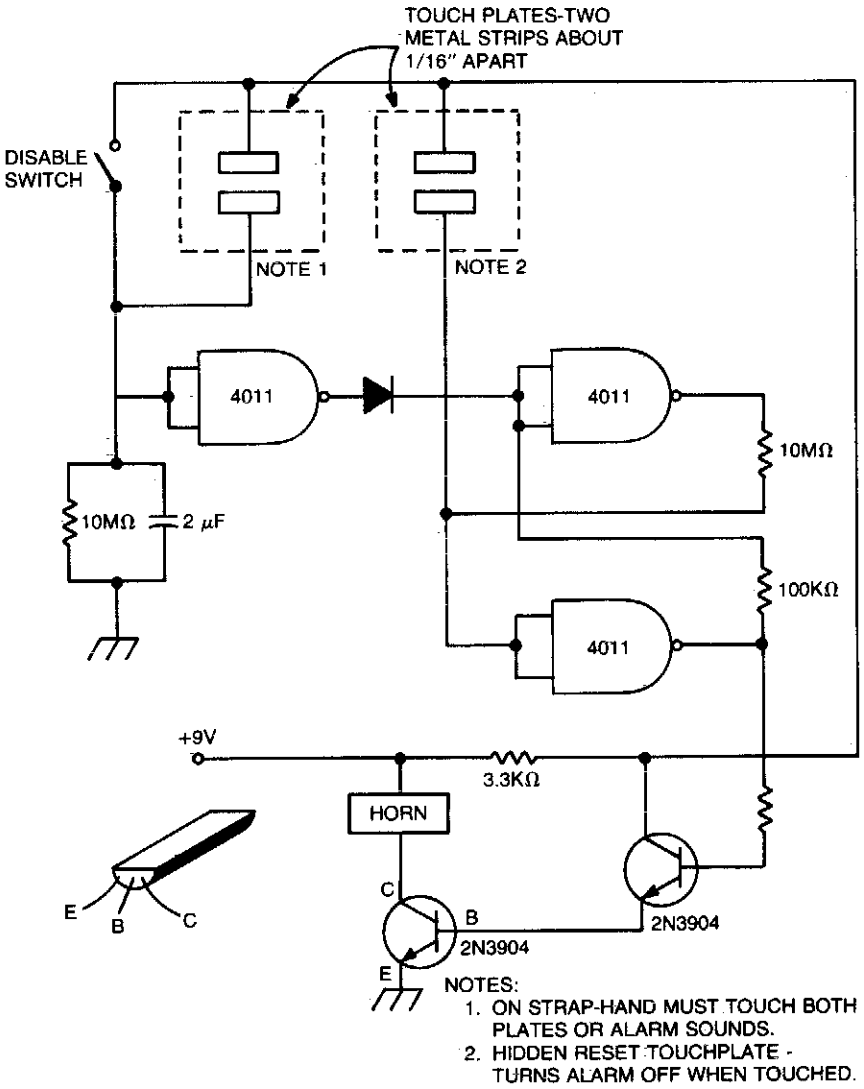
+9V
R3
470Ω
R2
3.3MΩ
D1
TIL209
T1
4
8
6
555
3
5
1
2
D2
TIL209
R1
3.3MΩ
C1
0.1μF
T2
R4
470Ω

Fig. 14-7

Circuit Notes

This circuit uses a 555 timer in the bistable mode. Touching T2 causes the output to go high; D2 conducts and D1 extinguishes. Touching T1 causes the output to go low; D1 conducts and D2 is cut off. The output from pin 3 can also be used to operate other circuits (e.g., a triac controlled lamp). In this case, the LEDs are useful for finding the touch terminals in the dark. C1 is not absolutely necessary but helps to prevent triggering from spurious pulses.

CAPACITANCE OPERATED ALARM TO FOIL PURSE SNATCHERS


TOUCH PLATES-TWO METAL STRIPS ABOUT 1/16" APART
DISABLE SWITCH
NOTE 1
NOTE 2
4011
4011
4011
10MΩ
10MΩ
2 µF
100KΩ
+9V
3.3KΩ
HORN
C
B
E
2N3904
2N3904
E
B
C

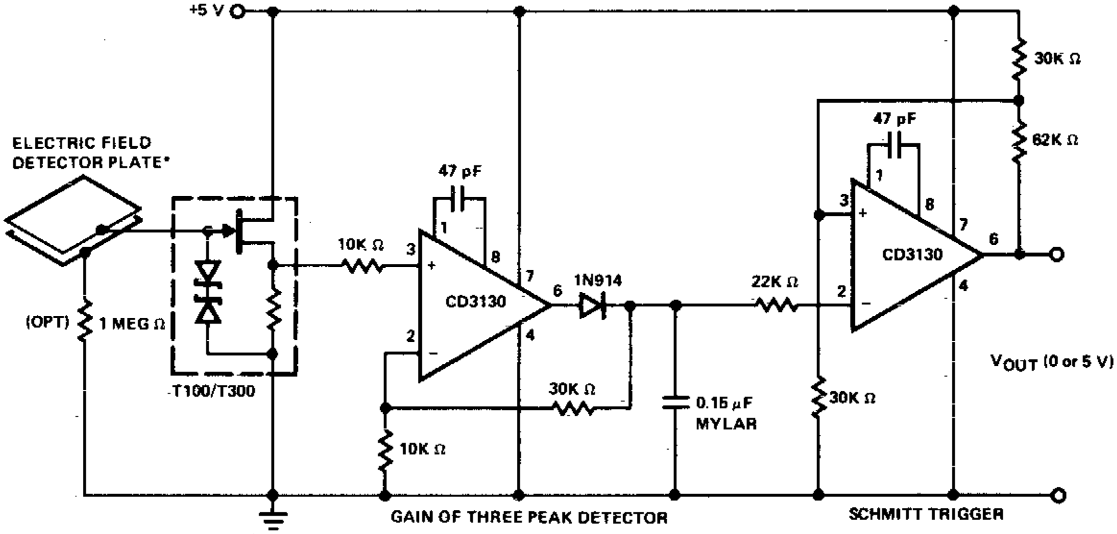
NOTES:
1. ON STRAP-HAND MUST TOUCH BOTH PLATES OR ALARM SOUNDS.
2. HIDDEN RESET TOUCHPLATE - TURNS ALARM OFF WHEN TOUCHED.

Fig. 14-8

Circuit Notes

As long as touch plates (1) are touched together, the alarm is off. If not held for about 30 seconds, the alarm goes off. The circuit can be disabled with switch or by touching the plates (2). The alarm is battery operated by a bicycle horn.

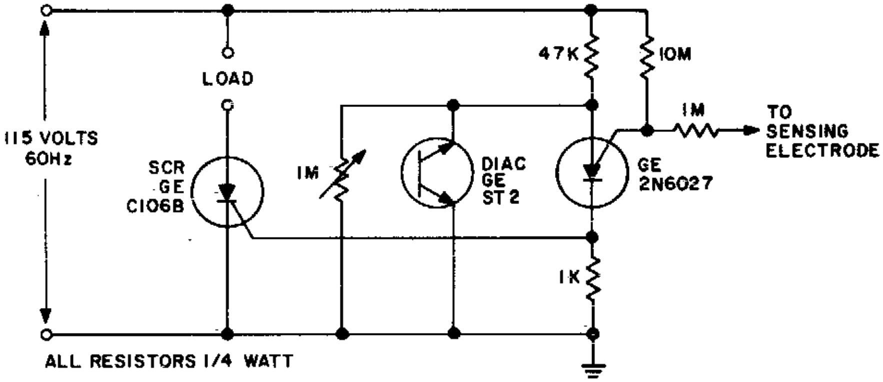
SELF-BIASED PROXIMITY SENSOR WORKS ON DETECTED CHANGING FIELD


+5 V
ELECTRIC FIELD DETECTOR PLATE*
(OPT)
1 MEG Ω
T100/T300
10K Ω
47 pF
CD3130
1N914
30K Ω
10K Ω
0.15 μF MYLAR
22K Ω
30K Ω
62K Ω
CD3130
VOUT (0 or 5 V)
GAIN OF THREE PEAK DETECTOR
SCHMITT TRIGGER

*DETECTOR PLATE MAY BE DOUBLE-SIDED PC BOARD OR ANY INSULATED METAL SHEET

Fig. 14-9

TOUCH SWITCH OR PROXIMITY DETECTOR


115 VOLTS 60Hz
LOAD
SCR GE C106B
IM
DIAC GE ST2
GE 2N6027
47K
10M
1M
TO SENSING ELECTRODE
1K

ALL RESISTORS 1/4 WATT

Fig. 14-10

Circuit Notes

This circuit is actuated by an increase in capacitance between a sensing electrode and the ground side of the line. The sensitivity can be adjusted to switch when a human body is within inches of the insulated plate used as the sensing electrode. Thus, sensitivity is adjusted with the 1 megohm potentiometer which determines the anode voltage level prior to clamping. This sensitivity will be proportional to the area of the surface opposing each other.

FINGER TOUCH OR CONTACT SWITCH

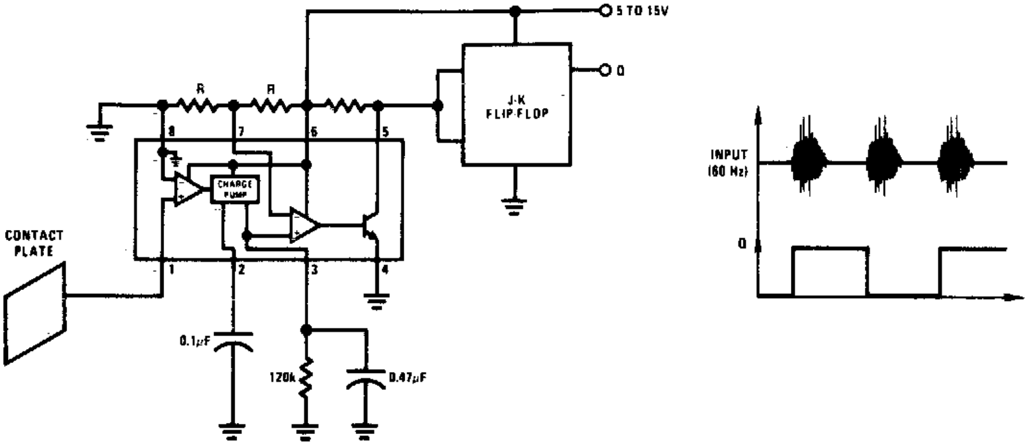

5 TO 15V
Q
J-K FLIP-FLOP
R
R
8
7
6
5
CHARGE PUMP
CONTACT PLATE
1
2
3
4
0.1µF
120k
0.47µF
INPUT (60 Hz)
0

Fig. 14-11

PROXIMITY DETECTOR

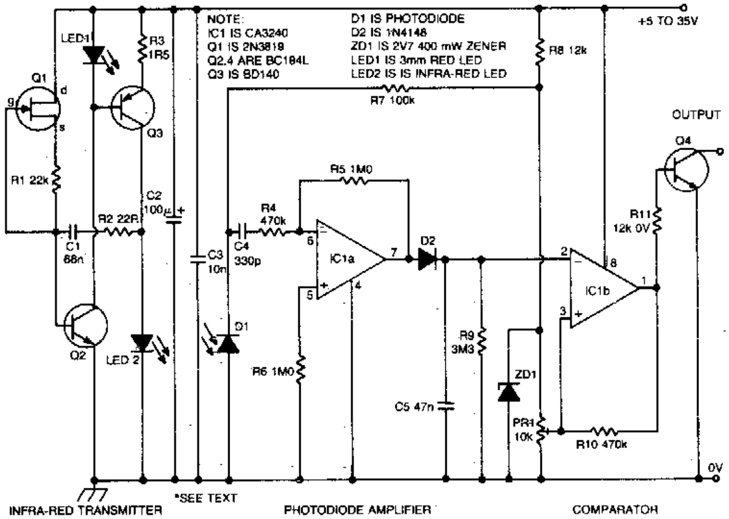

NOTE:
IC1 IS CA3240
Q1 IS 2N3819
Q2.4 ARE BC184L
Q3 IS BD140
D1 IS PHOTODIODE
D2 IS 1N4148
ZD1 IS 2V7 400 mW ZENER
LED1 IS 3mm RED LED
LED2 IS IS INFRA-RED LED
+5 TO 35V
R8 12k
R7 100k
LED1
R3 1R5
Q1
g
d
s
Q3
R1 22k
R2 22R
C2 100µ
C1 68n
C3 10n
C4 330p
R4 470k
R5 1M0
IC1a
6
5
7
4
D2
D1
R6 1M0
R9 3M3
C5 47n
ZD1
PR1 10k
R10 470k
IC1b
2
3
8
1
R11 12k 0V
Q4
OUTPUT
0V
Q2
LED 2
*SEE TEXT
INFRA-RED TRANSMITTER
PHOTODIODE AMPLIFIER
COMPARATOR

Fig. 14-12

Circuit Notes

The proximity sensor works on the principle of transmitting a beam of modulated infra-red light from the emitter diode LED2, and receiving reflections from objects passing in front of the beam with a photodiode detector D1. The circuit can be split into three distinct stages; the infra-red transmitter, the photodiode amplifier, and a variable threshold comparator.

TOUCH CIRCUIT

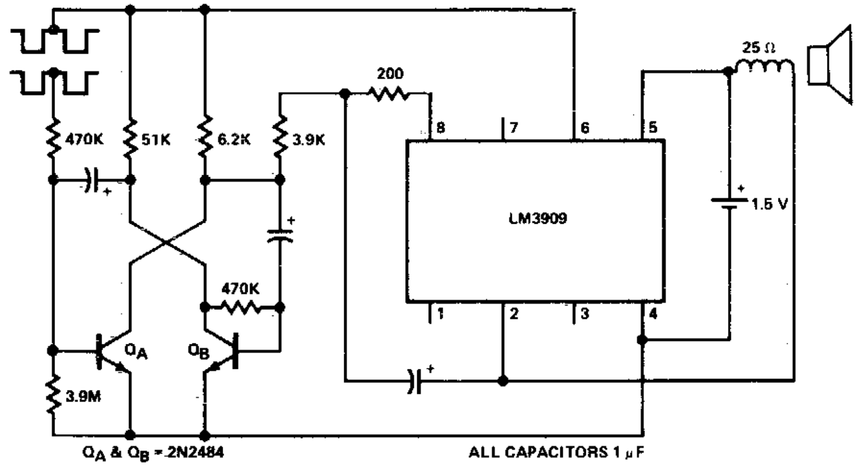

200
25 Ω
470K
51K
6.2K
3.9K
8
7
6
5
LM3909
1.5 V
470K
1
2
3
4
Q_A
Q_B
3.9M
Q_A & Q_B = 2N2484
ALL CAPACITORS 1 μF

Fig. 14-13

CMOS TOUCH SWITCH

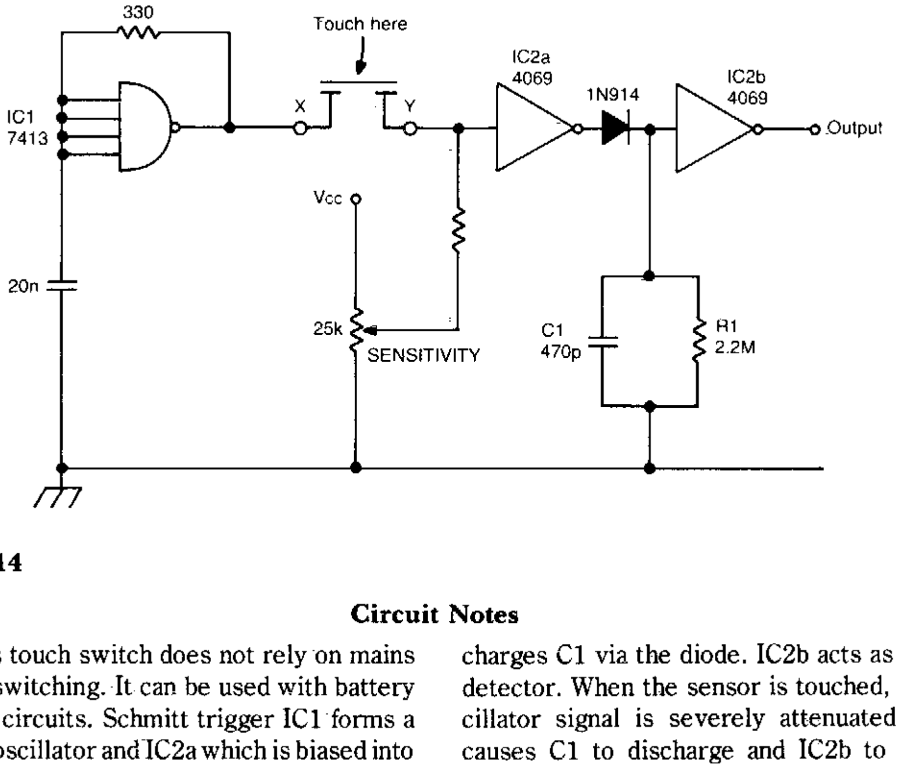

330
Touch here
IC1
7413
X
Y
IC2a
4069
1N914
IC2b
4069
Output
Vcc
20n
25k
SENSITIVITY
C1
470p
R1
2.2M

Fig. 14-14

Circuit Notes

This touch switch does not rely on mains hum for switching. It can be used with battery powered circuits. Schmitt trigger IC1 forms a 100 kHz oscillator and IC2a which is biased into the linear region, amplifies the output and charges C1 via the diode. IC2b acts as a level detector. When the sensor is touched, the oscillator signal is severely attenuated which causes C1 to discharge and IC2b to change state.

LATCHING, DOUBLE BUTTON TOUCH SWITCH

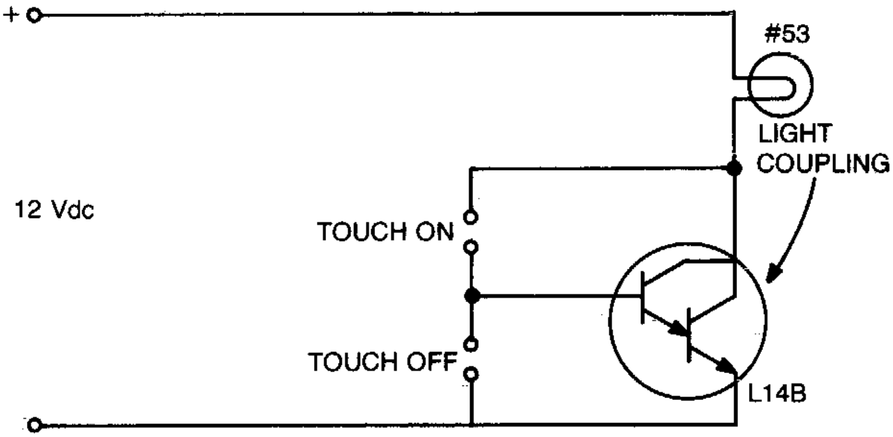

+
12 Vdc
#53
LIGHT COUPLING
TOUCH ON
TOUCH OFF
L14B

Fig. 14-15

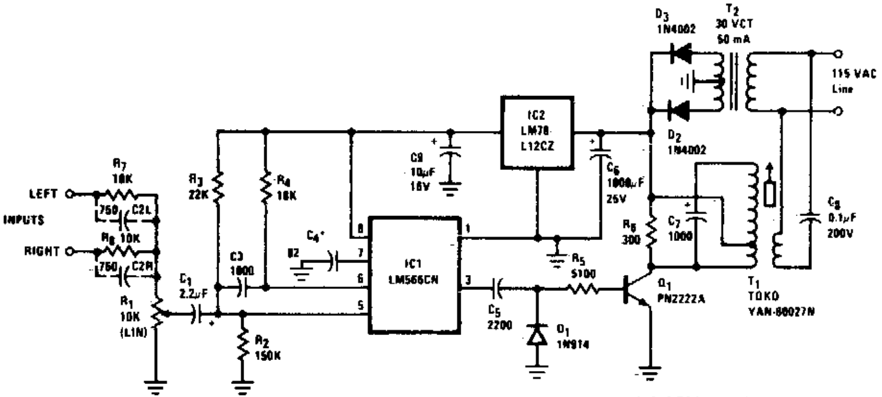
15

Carrier Current Circuits

The sources of the following circuits are contained in the Sources section beginning on page 730. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

FM Carrier Current Remote Speaker System
200 kHz Line Carrier Transmitter with On/Off Carrier Modulation
Carrier Current Receiver
Carrier Current Transmitter
Carrier Current Transmitter
Integrated Circuit Current Transmitter
Single Transistor Carrier Current Receiver
IC Carrier-Current Receiver
Carrier-Current Remote Control or Intercom

Carrier System Transmitter


LEFT
INPUTS
RIGHT
R_7 10K
750
C2L
R_8 10K
750
C2R
R_1 10K (LIN)
C_1 2.2μF
+
R_3 22K
R_4 10K
C3 1000
R_2 150K
C_4*
82
IC1 LM566CN
8
7
6
5
1
3
C9 10μF 16V
+
IC2 LM78L12CZ
C_6 1000μF 25V
+
R_5 5100
C_5 2200
D_1 1N914
Q_1 PN2222A
R_6 300
C_7 1000
+
D_3 1N4002
D_2 1N4002
T_2 30 VCT 50 mA
115 VAC Line
C_8 0.1μF 200V
T_1 TOKO YAN-80027N

CAPACITOR VALUES IN pF
RESISTOR VALUES IN Ω
†SELECT FOR CARRIER FREQ.

f_c	C_4	C_7
200 kHz	82	1000
100 kHz	160	3900

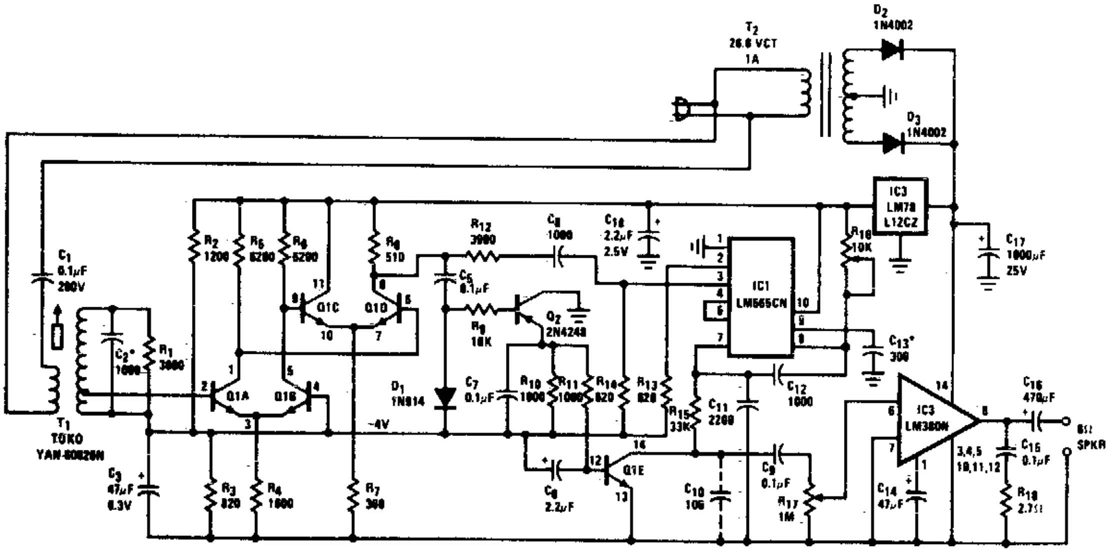
Carrier System Receiver

FM CARRIER CURRENT REMOTE SPEAKER SYSTEM

Circuit Notes

High quality, noise free, wireless FM transmitter/receiver operates over standard power lines. Complete system is suitable for high-quality transmission of speech or music, and will operate from any ac outlet anywhere on a one-acre homesite. Frequency response is 20-20, 000 Hz and THD is under ½%. Transmission distance along a power line is at least adequate to include all outlets in and around a suburban home and yard.

Two input terminals are provided so that both left and right signals of a stereo set may be combined for mono transmission to a single remote speaker if desired.


T2
26.8 VCT
1A
D2
1N4002
D3
1N4002
IC3
LM78
L12CZ
C17
1000µF
25V
C1
0.1µF
200V
C2*
1000
R1
3000
T1
TOKO
YAN-60826N
C3
47µF
6.3V
R2
1200
R5
6200
R6
6200
R8
510
Q1A
Q1B
Q1C
Q1D
R3
820
R4
1000
R7
360
-4V
D1
1N914
R12
3900
C8
1000
C18
2.2µF
2.5V
C5
0.1µF
R9
10K
Q2
2N4248
C7
0.1µF
R10
1000
R11
1000
R14
820
R13
820
R15
33K
C6
2.2µF
Q1E
C10
106
C11
2200
C12
1000
C9
0.1µF
R17
1M
IC1
LM565CN
R16
10K
C13*
300
IC3
LM380N
3,4,5
10,11,12
C14
47µF
C16
470µF
C15
0.1µF
R19
2.7Ω
8Ω
SPKR

CAPACITOR VALUES IN PF
RESISTOR VALUES IN Ω
*SELECT FOR CARRIER FREQ.

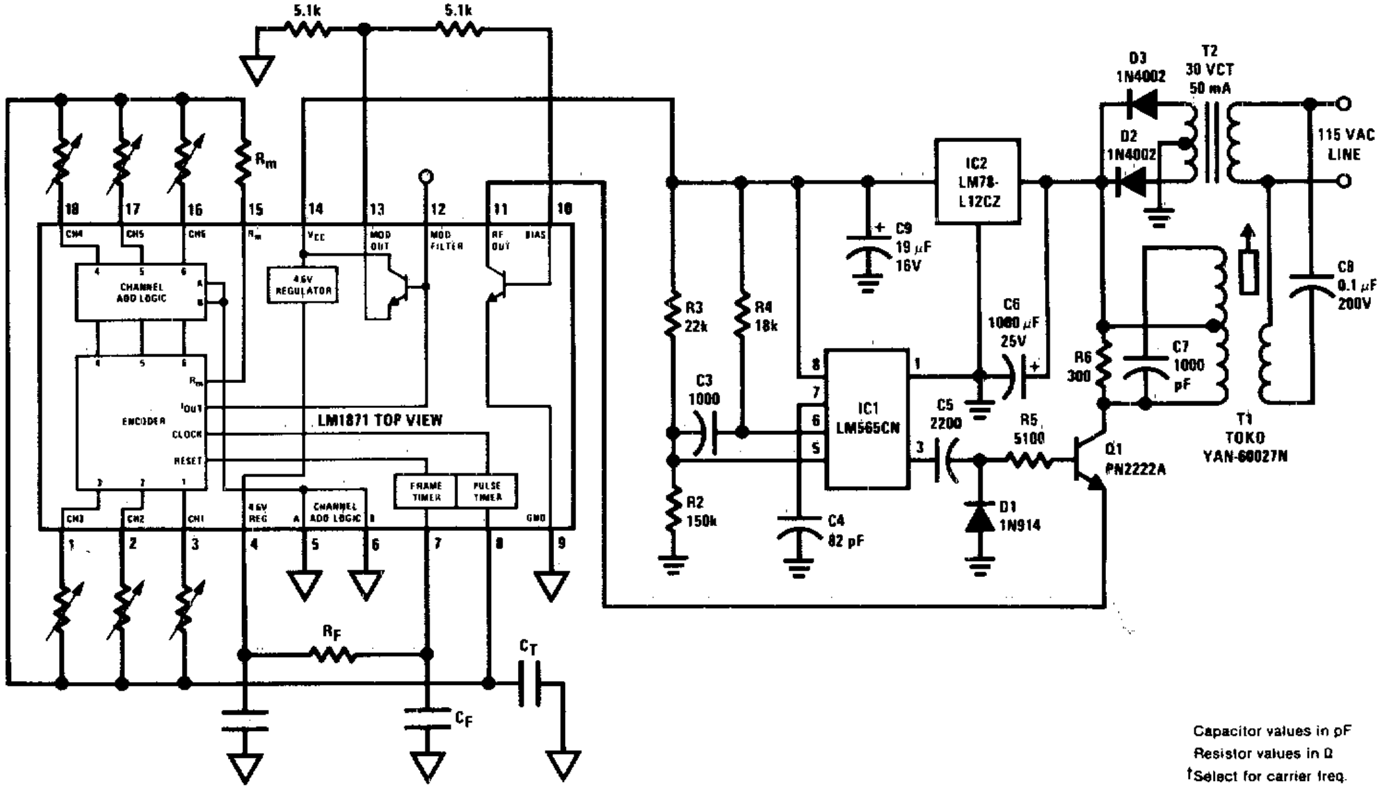
Q1A–Q1E = LM3046

f_0	C_2	C_{13}
200 kHz	1000	300
100 kHz	3000	620

Fig. 15-1

The receiver amplifies, limits, and demodulates the received FM signal. It provides audio mute in the absence of carrier and 2.5 W output to a speaker.

200 kHz LINE CARRIER
TRANSMITTER WITH ON/OFF CARRIER MODULATOR

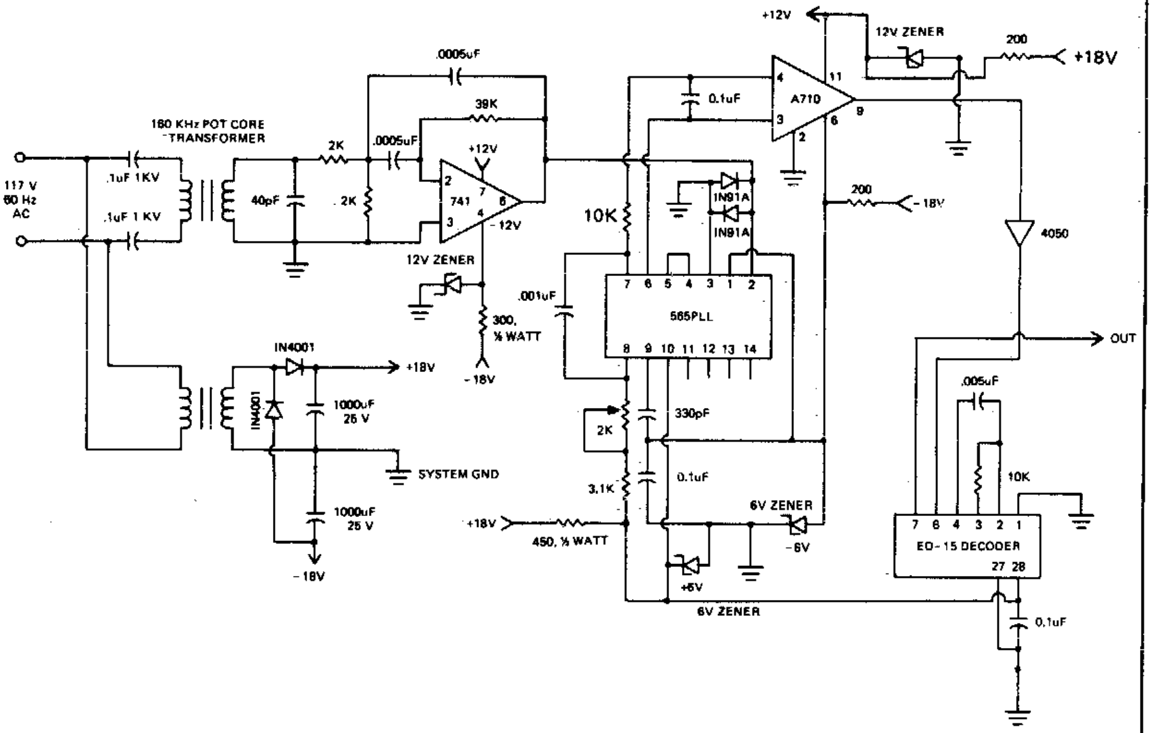

5.1k
5.1k
D3
1N4002
T2
30 VCT
50 mA
D2
1N4002
115 VAC
LINE
IC2
LM78-
L12CZ
C9
19 µF
16V
C8
0.1 µF
200V
C6
1000 µF
25V
R3
22k
R4
18k
R6
300
C7
1000
pF
C3
1000
IC1
LM565CN
C5
2200
R5
5100
Q1
PN2222A
T1
TOKO
YAN-60027N
R2
150k
C4
82 pF
D1
1N914
R_m
CH4
CH5
CH6
R_m
V_{CC}
MOD OUT
MOD FILTER
RF OUT
BIAS
CHANNEL ADD LOGIC
4.6V REGULATOR
ENCODER
I_{OUT}
CLOCK
RESET
LM1871 TOP VIEW
FRAME TIMER
PULSE TIMER
CH3
CH2
CH1
4.6V REG
CHANNEL ADD LOGIC
GND
R_F
C_T
C_F
1
2
3
4
5
6
7
8
9
10
11
12
13
14
15
16
17
18

Capacitor values in pF
Resistor values in Ω
†Select for carrier freq.

f_c	C4	C7
200 kHz	82	1000
100 kHz	160	3900

Fig. 15-2

CARRIER CURRENT RECEIVER



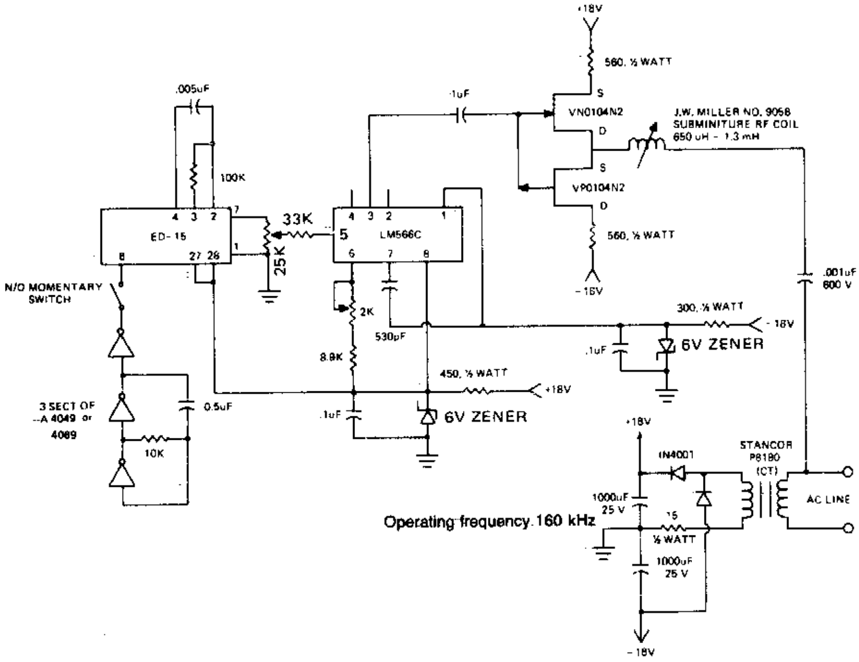
117 V
60 Hz
AC
.1uF 1KV
.1uF 1 KV
160 KHz POT CORE TRANSFORMER
40pF
2K
.2K
.0005uF
.0005uF
39K
+12V
741
-12V
12V ZENER
300, ½ WATT
-18V
.001uF
IN4001
+18V
1000uF 25 V
SYSTEM GND
1000uF 25 V
-18V
10K
IN91A
IN91A
565PLL
7 6 5 4 3 1 2
8 9 10 11 12 13 14
330pF
2K
3.1K
0.1uF
6V ZENER
-6V
+18V
450, ½ WATT
+6V
6V ZENER
+12V
0.1uF
A710
12V ZENER
200
+18V
200
-18V
4050
OUT
.005uF
10K
7 6 4 3 2 1
EO-15 DECODER
27 28
0.1uF

Fig. 15-3

Circuit Notes

160 kHz transformer consists of a 18 × 11mm ungapped pot core (Siemens, Ferrocube, etc.), utilizing magnetics incorporated type "F" material wound with 80½ turns of No. 35 wire for the secondary and 5½ turns for the primary. This gives a turns ratio of approximately 15 to 1.

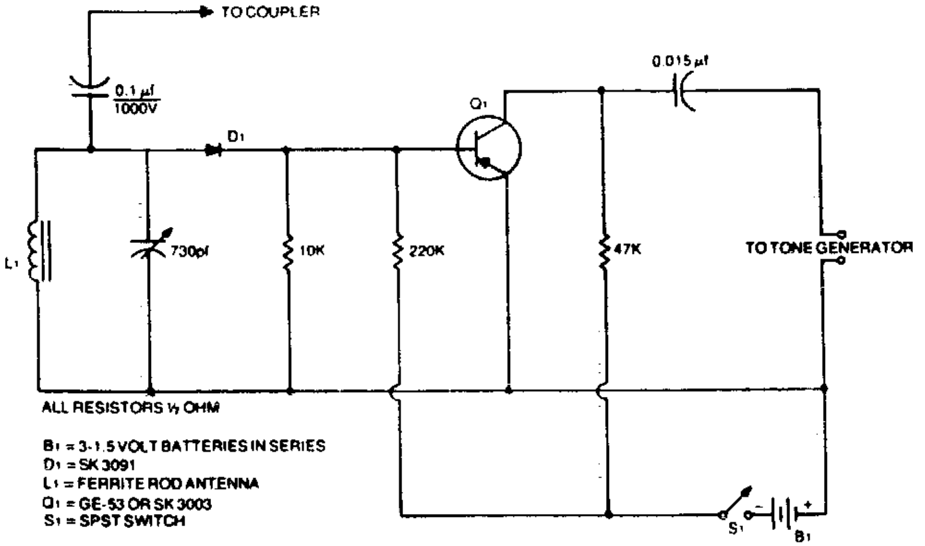
CARRIER CURRENT TRANSMITTER


+18V
560, ½ WATT
S
.1uF
VN0104N2
D
J.W. MILLER NO. 9058
SUBMINITURE RF COIL
650 uH - 1.3 mH
S
VP0104N2
D
560, ½ WATT
-18V
.005uF
100K
4 3 2
7
ED-15
8
27 28
1
33K
25K
4 3 2
1
5
LM566C
6 7 8
N/O MOMENTARY
SWITCH
2K
530pF
8.9K
.001uF
600 V
300, ½ WATT
-18V
.1uF
6V ZENER
450, ½ WATT
+18V
3 SECT OF
--A 4049 or
4069
0.5uF
10K
.1uF
6V ZENER
+18V
IN4001
STANCOR
P8180
(CT)
AC LINE
1000uF
25 V
15
½ WATT
1000uF
25 V
-18V

Operating frequency 160 kHz

Fig. 15-4

CARRIER CURRENT TRANSMITTER

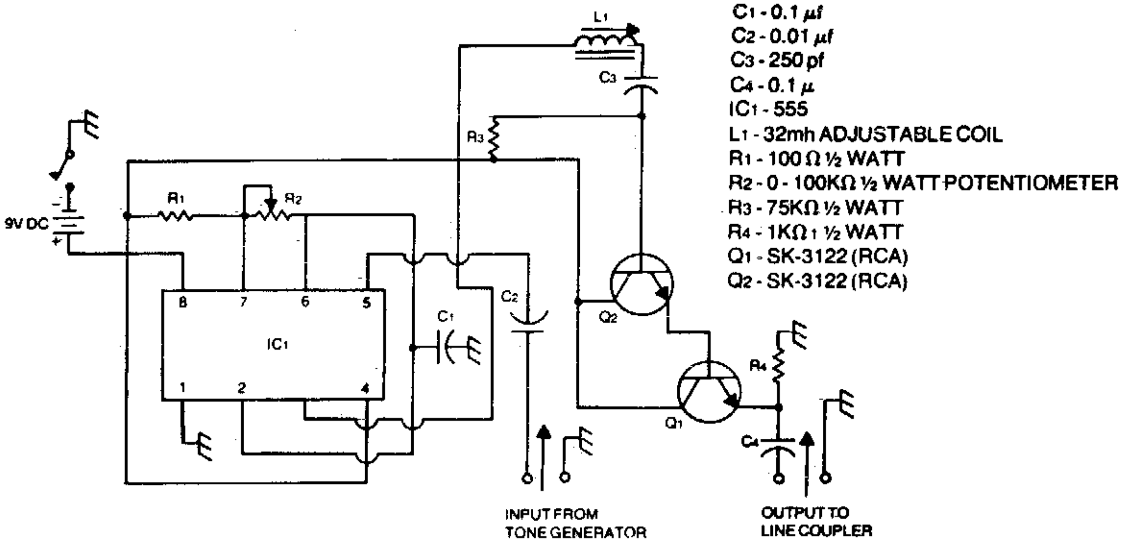

TO COUPLER
0.1 µf
1000V
D₁
Q₁
0.015 µf
L₁
730pf
10K
220K
47K
TO TONE GENERATOR
S₁
B₁
-
+

ALL RESISTORS ½ OHM

B₁ = 3-1.5 VOLT BATTERIES IN SERIES
D₁ = SK 3091
L₁ = FERRITE ROD ANTENNA
Q₁ = GE-53 OR SK 3003
S₁ = SPST SWITCH

Fig. 15-5

IC CARRIER CURRENT TRANSMITTER


9V DC
R_1
R_2
R_3
R_4
L_1
C_1
C_2
C_3
C_4
IC_1
8
7
6
5
1
2
4
Q_1
Q_2
INPUT FROM TONE GENERATOR
OUTPUT TO LINE COUPLER

C_1 - 0.1 µf
C_2 - 0.01 µf
C_3 - 250 pf
C_4 - 0.1 µ
IC_1 - 555
L_1 - 32mh ADJUSTABLE COIL
R_1 - 100 Ω ½ WATT
R_2 - 0 - 100KΩ ½ WATT POTENTIOMETER
R_3 - 75KΩ ½ WATT
R_4 - 1KΩ 1 ½ WATT
Q_1 - SK-3122 (RCA)
Q_2 - SK-3122 (RCA)

Fig. 15-6

SINGLE TRANSISTOR CARRIER CURRENT RECEIVER

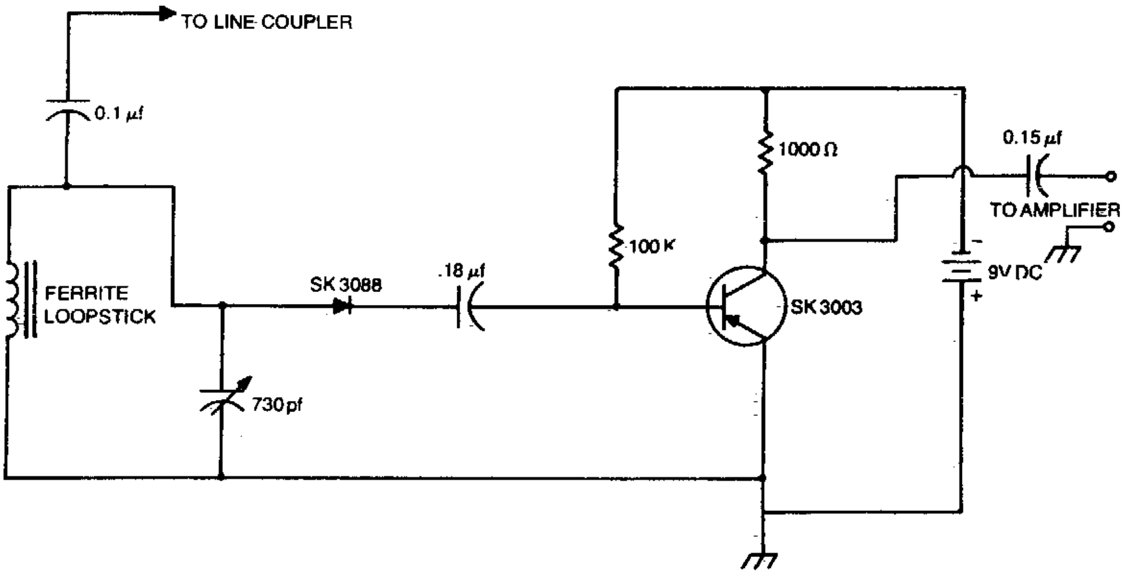
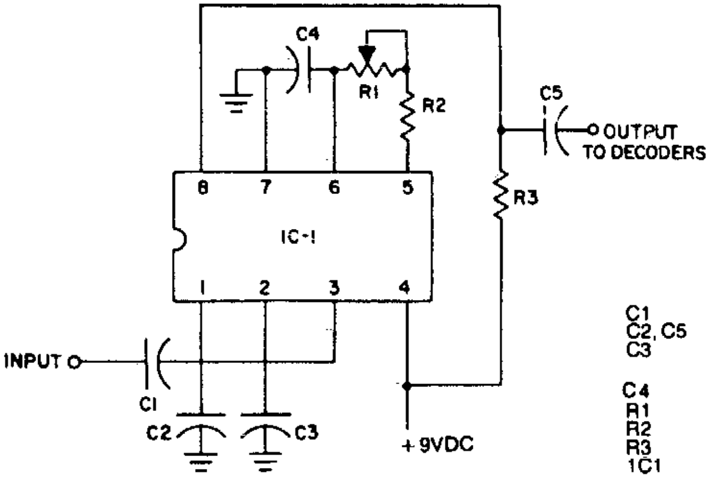

TO LINE COUPLER
0.1 µf
FERRITE LOOPSTICK
730 pf
SK 3088
.18 µf
100 K
1000 Ω
SK 3003
0.15 µf
TO AMPLIFIER
9V DC

Fig. 15-7

IC CARRIER-CURRENT RECEIVER


C4
R1
R2
C5
OUTPUT
TO DECODERS
8
7
6
5
IC-1
1
2
3
4
R3
INPUT
C1
C2
C3
+9VDC

C1	0.1 µF, 600V PAPER CAPACITOR
C2, C5	0.1 µF, 50V MYLAR CAPACITOR
C3	22µF, 35 VOLT ELECTROLYTIC CAPACITOR
C4	.0047 µF25 VOLT CAPACITOR
R1	20K POTENTIOMETER
R2	2K, ½ WATT RESISTOR
R3	4K, ½ WATT RESISTOR
1C1	TYPE 567 TONE DECODER

Fig. 15-8

CARRIER-CURRENT REMOTE CONTROL OR INTERCOM

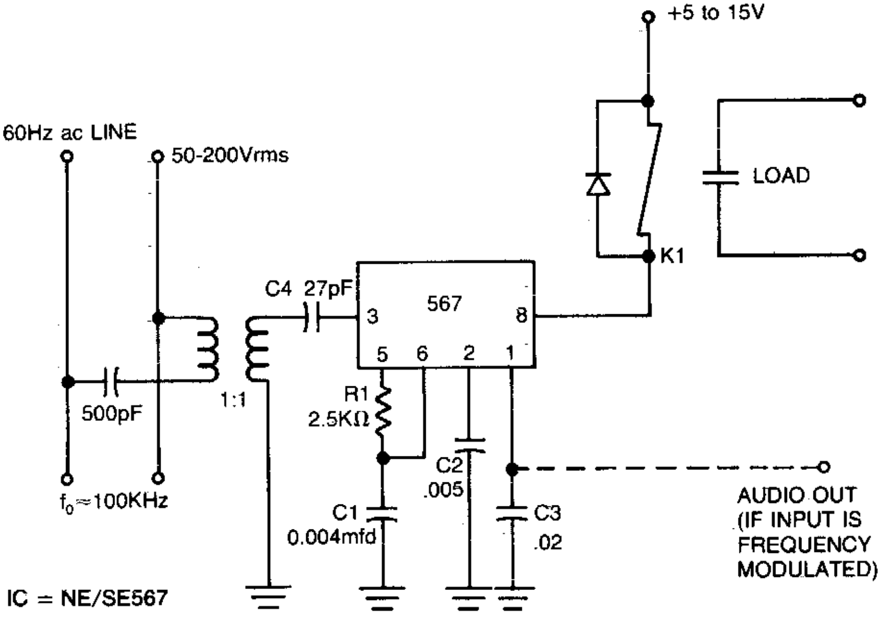

+5 to 15V
60Hz ac LINE
50-200Vrms
LOAD
K1
C4 27pF
567
3
8
5
6
2
1
500pF
1:1
R1
2.5KΩ
C2
.005
C1
0.004mfd
C3
.02
f₀≈100KHz
AUDIO OUT
(IF INPUT IS
FREQUENCY
MODULATED)
IC = NE/SE567

Fig. 15-9

16

Comparators

The sources of the following circuits are contained in the Sources section beginning on page 730. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

Null Detector
Comparator with Variable Hysteresis
Diode Feedback Comparator
Undervoltage/Overvoltage Indicator
Dual Limit Comparator
High/Low Limit Alarm
Window Comparator
Window Comparator Driving High/Low Lamps
Comparator with Time Out
Noninverting Comparator with Hysteresis
Inverting Comparator with Hysteresis
Window Comparator
Micropower Double-Ended Limit Detector
Opposite Polarity Input Voltage Comparator
Limit Comparator
Comparator Clock Circuit
Double-Ended Limit Comparator
Limit Comparator
Precision, Dual Limit Go/No Go Tester
Comparator with Hysteresis
High Impedance Comparator
Comparator

NULL DETECTOR

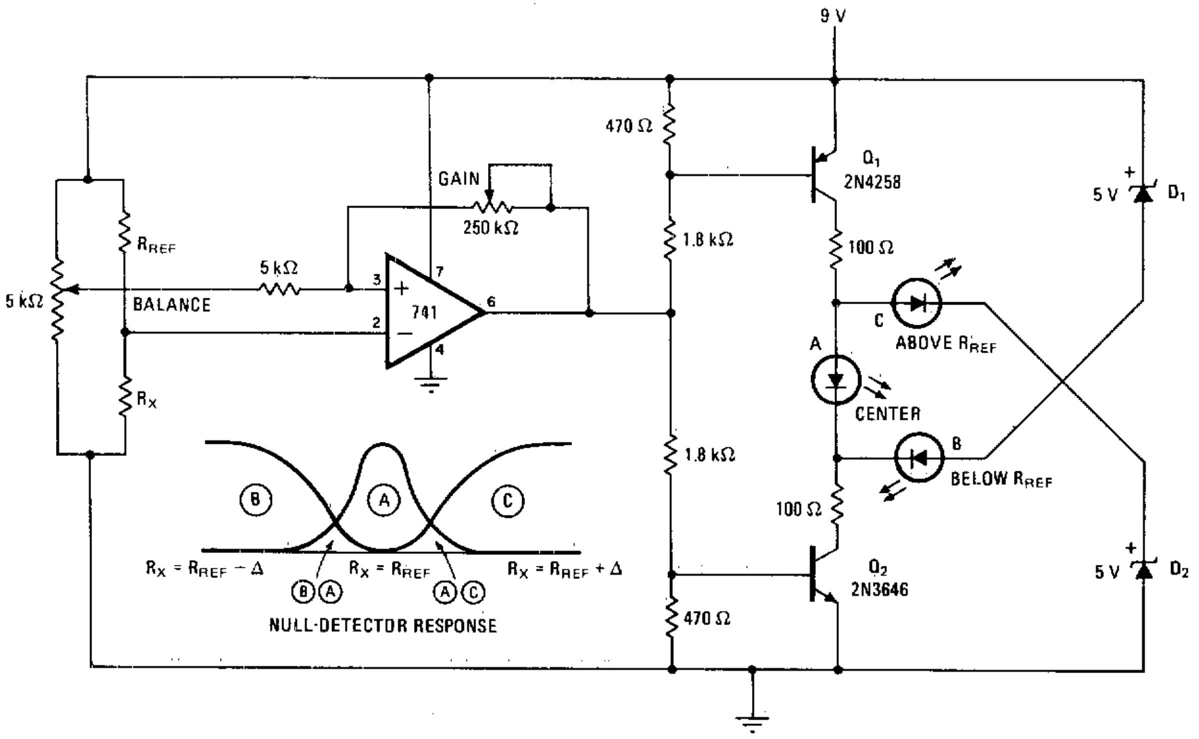

9 V
470 Ω
Q₁
2N4258
5 V
+
D₁
GAIN
250 kΩ
R_REF
5 kΩ
1.8 kΩ
100 Ω
5 kΩ
BALANCE
3
+
741
7
6
2
−
4
C
ABOVE R_REF
A
CENTER
R_X
B
BELOW R_REF
B
A
C
1.8 kΩ
100 Ω
R_X = R_REF − Δ
R_X = R_REF
R_X = R_REF + Δ
B A
A C
NULL-DETECTOR RESPONSE
Q₂
2N3646
5 V
+
D₂
470 Ω

Fig. 16-1

Circuit Notes

Null detector uses simple LED readout to indicate if test resistor R_x is below, equal to, or greater than test resistance R_{ref}. If $R_x = R_{ref}$, the 741 output sits at midpoint value of 4.5 volts and LED A lights. Otherwise, the output of the 741 turns off one transistor and diverts current from the other transistor through B or C, depending on the polarity of the input voltage difference. Null-detector response is illustrated.

COMPARATOR WITH VARIABLE HYSTERESIS (WITHOUT SHIFTING INITIAL TRIP POINT)

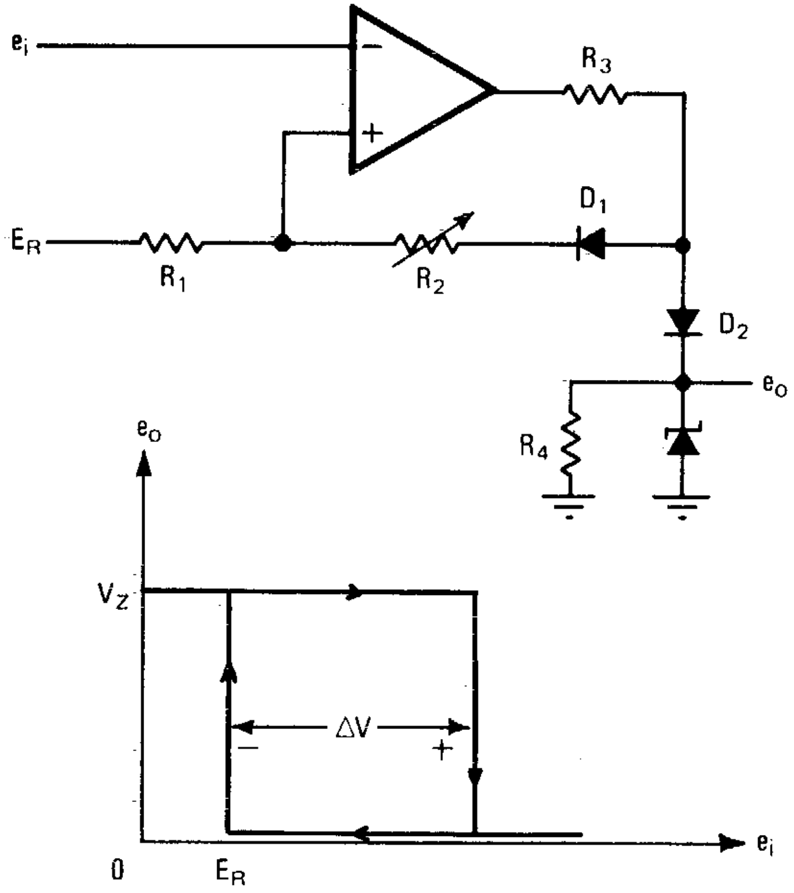
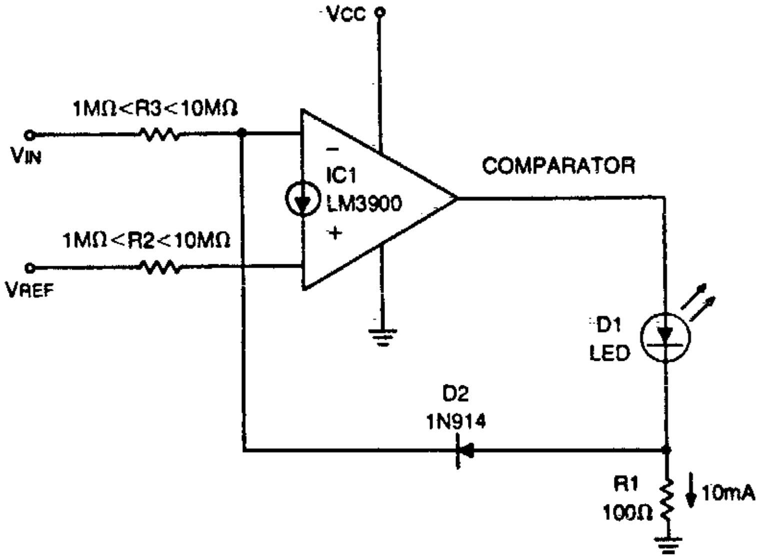


Fig. 16-2

Circuit Notes

An operational amplifier can be used as a convenient device for analog comparator applications that require two different trip points. The addition of a positive-feedback network introduces a precise variable hysteresis into the usual comparator switching action. Such feedback develops two comparator trip points centered about the initial trip point or reference point. The voltage difference, ΔV, between the trip points can be adjusted by varying resistor R2. When the output voltage is taken from the zener diode, as shown, it switches between zero and Vz, the zener voltage.

DIODE FEEDBACK COMPARATOR


VCC
1MΩ<R3<10MΩ
VIN
IC1
LM3900
−
+
COMPARATOR
1MΩ<R2<10MΩ
VREF
D1
LED
D2
1N914
R1
100Ω
10mA

Parts list
IC1—LM3900
D1—LED Lafayette 32P06331V
D2—IN914
All resistors 1/4W
R1—100Ω
R2—(See circuit) 1MΩ to 10MΩ
R3—(See circuit) 1MΩ to 10MΩ

Fig. 16-3

Circuit Notes

This circuit can drive an LED display with constant current independently of wide power supply voltage changes. It can operate with a power supply range of at least 4V to 30V. With 10M resistances for R2 and R3 and the inverting input of the comparator grounded, the circuit becomes an LED driver with very high input impedance. The circuit can also be used in many other applications where a controllable constant current source is needed.

UNDERVOLTAGE/OVERVOLTAGE INDICATOR

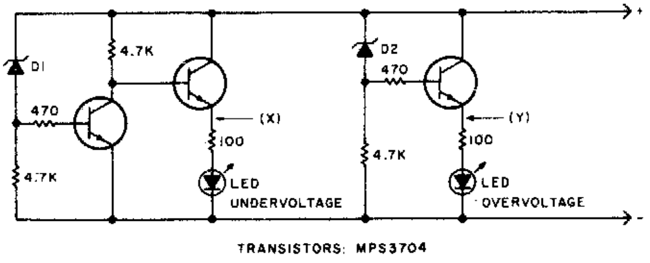

+
D1
4.7K
470
4.7K
(X)
100
LED
UNDERVOLTAGE
D2
470
4.7K
(Y)
100
LED
OVERVOLTAGE
−
TRANSISTORS: MPS3704

Fig. 16-4

Circuit Notes

This circuit will make the appropriate LED glow if the monitored voltage goes below or above the value determined by zener diodes D1 and D2.

DUAL LIMIT COMPARATOR

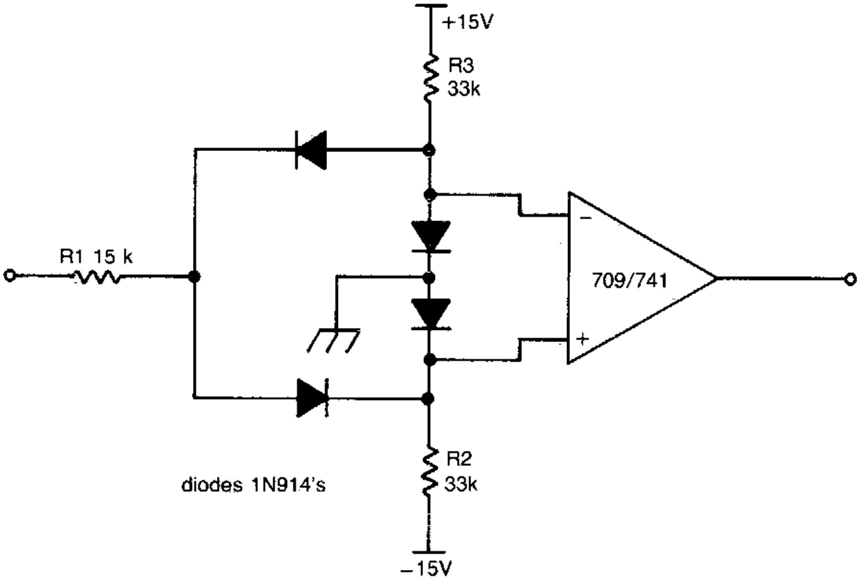
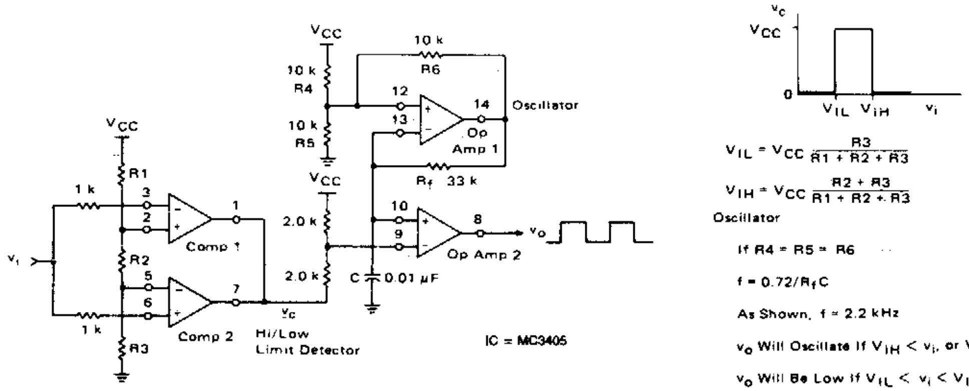

+15V
R3
33k
R1 15 k
709/741
−
+
diodes 1N914's
R2
33k
−15V

Fig. 16-5

Circuit Notes

This circuit gives a positive output when the input voltage exceeds 8.5 volts. Between these limits the output is negative. The positive limit point is determined by the ratio of R1, R2, and the negative point by R1, R3. The forward voltage drop across the diodes must be allowed for. The output may be inverted by reversing the inputs to the op amp. The 709 is used without frequency compensation.

HIGH/LOW LIMIT ALARM


V_{CC}
10 k
R4
10 k
R5
10 k
R6
12
13
14
Oscillator
Op Amp 1
R_f 33 k
V_{CC}
R1
1 k
3
2
1
Comp 1
v_i
R2
5
6
7
1 k
R3
Comp 2
V_{CC}
2.0 k
2.0 k
C 0.01 μF
10
9
8
v_o
Op Amp 2
v_c
Hi/Low Limit Detector
IC = MC3405
v_c
V_{CC}
0
V_{IL}
V_{IH}
v_i

$$V_{IL} = V_{CC} \frac{R3}{R1 + R2 + R3}$$

$$V_{IH} = V_{CC} \frac{R2 + R3}{R1 + R2 + R3}$$

Oscillator

If $R4 = R5 = R6$

$f = 0.72/R_f C$

As Shown, $f = 2.2$ kHz

v_o Will Oscillate If $V_{IH} < v_i$, or $V_{IL} > v_i$

v_o Will Be Low If $V_{IL} < v_i < V_{IH}$

Fig. 16-6

WINDOW COMPARATOR

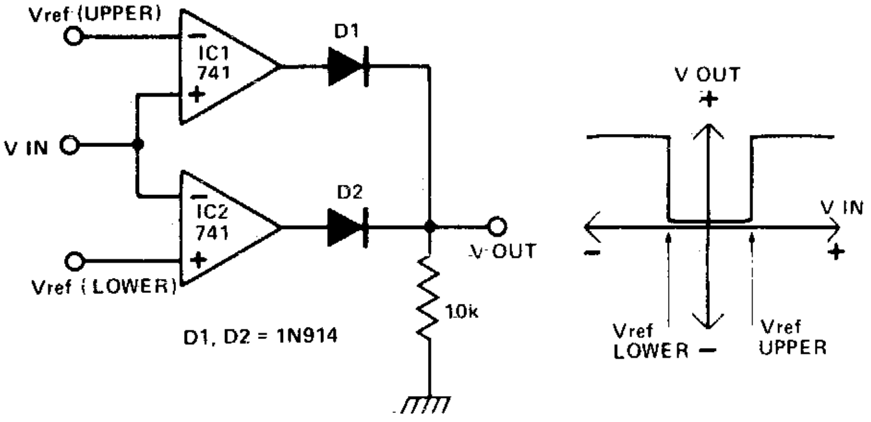
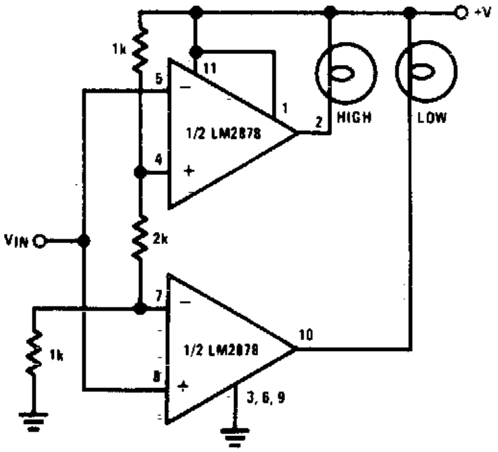

Vref (UPPER)
IC1
741
D1
V IN
IC2
741
D2
Vref (LOWER)
V OUT
10k
D1, D2 = 1N914
V OUT
+
V IN
−
+
Vref
LOWER −
Vref
UPPER

Fig. 16-7

Circuit Notes

This circuit gives an output (which in this case is 0V) when an input voltage lies in between two specified voltages. When it is outside this window, the output is positive. The two op amps are used as voltage comparators. When Vin is more positive than Vref (upper) the output of IC1 is positive and D1 is forward biased. Otherwise the output is negative, D1 reverse biased and hence Vout is 0V. Similarly, when Vin is more negative than Vref (lower), the output of IC2 is positive; D2 is forward biased and this Vout is positive. Otherwise Vout is 0V. When Vin lies within the window set by the reference voltages, Vout is 0V.

WINDOW COMPARATOR DRIVING HIGH/LOW LAMPS


+V
1k
11
5
1
2
HIGH
LOW
1/2 LM2878
4
2k
VIN
7
10
1k
8
1/2 LM2878
3, 6, 9

TRUTH TABLE

V_{IN}	High	Low
< 1/4 V+	Off	On
1/4 V+ to 3/4 V+	Off	Off
> 3/4 V+	On	Off

Fig. 16-8

COMPARATOR WITH TIME OUT

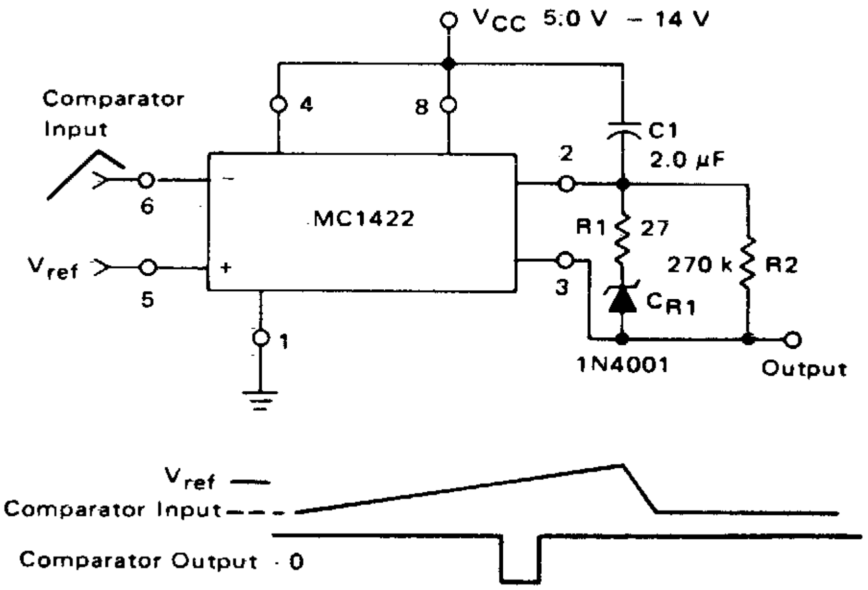
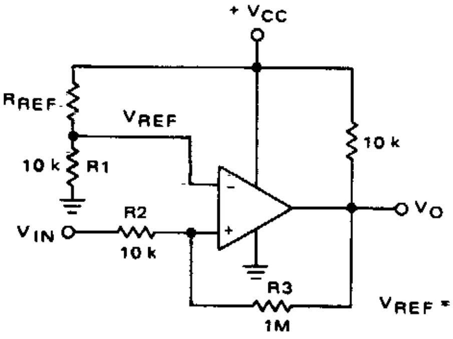

Vcc 5.0 V – 14 V
Comparator Input
4
8
C1 2.0 μF
2
6
MC1422
R1 27
270 k R2
Vref
5
3
CR1
1
1N4001
Output
Vref
Comparator Input
Comparator Output 0

Fig. 16-9

Circuit Notes

The MC1422 is used as a comparator with the capability of a timing output pulse when the inverting input (Pin 6) is ≥ the noninverting input (Pin 5). The frequency of the pulses for the values of R2 and C1 as shown is approximately 2.0 Hz, and the pulse width 0.3 ms.

NONINVERTING COMPARATOR WITH HYSTERESIS


+Vcc
RREF
VREF
10 k
10 k R1
R2
VIN
10 k
Vo
R3
1M

IC = LM358

$V_{REF} = \frac{V_{CC}\,R1}{R_{REF} + R1}$

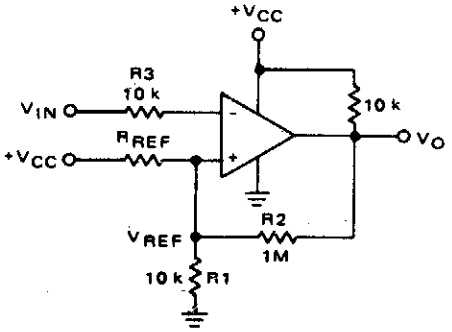
$R2 \approx R1 // R_{REF}$

Amount of Hysteresis V_M

$V_H = \frac{R2}{R2 + R3}\,(V_{Omax} - V_{Omin})$

Fig. 16-10

INVERTING COMPARATOR WITH HYSTERESIS


+VCC
R3
10 k
VIN
RREF
+VCC
10 k
VO
R2
VREF
1M
10 k
R1

$$V_{REF} \approx \frac{V_{CC}\,R1}{R_{REF} + R1}$$

$$R3 \approx R1 \,//\, R_{REF} \,//\, R1$$

$$V_H = \frac{R1//R_{REF}}{R1//R_{REF} + R2} \quad (V_{Omax} - V_{Omin})$$

Fig. 16-11

WINDOW COMPARATOR

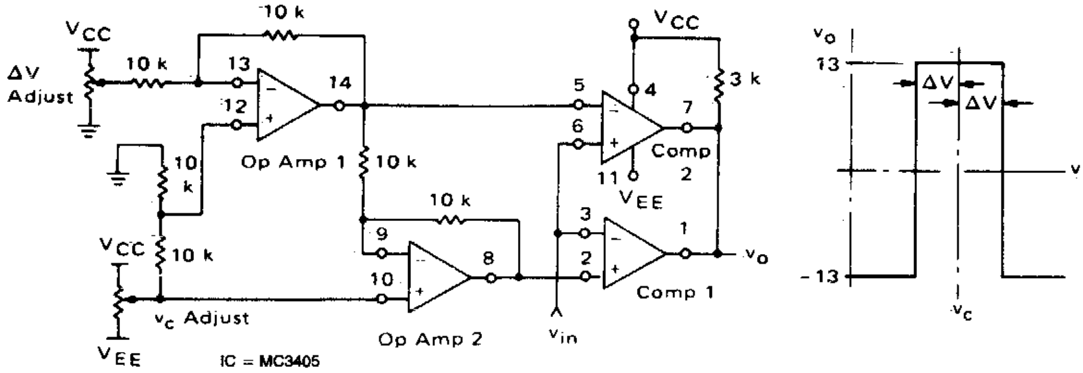

VCC
ΔV
Adjust
10 k
10 k
13
12
14
Op Amp 1
10
k
10 k
10 k
VCC
10 k
vc Adjust
VEE
IC = MC3405
10 k
9
10
8
Op Amp 2
vin
VCC
5
6
4
7
3 k
Comp
2
11
VEE
3
2
1
Comp 1
vo
vo
13
ΔV
ΔV
vin
- 13
vc

Fig. 16-12

MICROPOWER DOUBLE-ENDED LIMIT DETECTOR

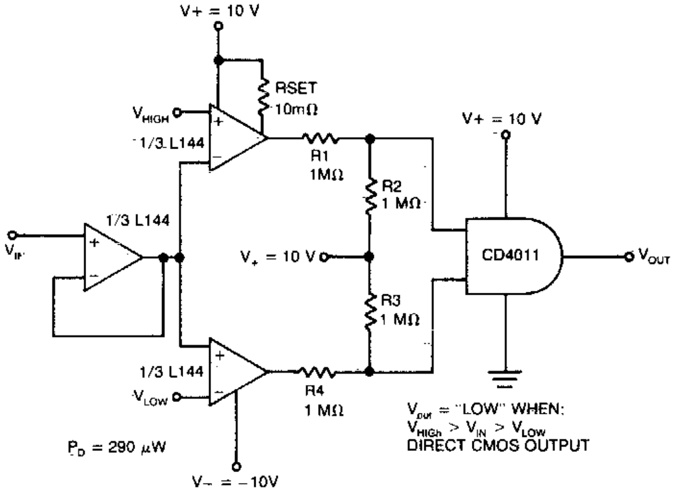

V+ = 10 V
RSET
10mΩ
VHIGH
1/3 L144
R1
1MΩ
V+ = 10 V
R2
1 MΩ
1/3 L144
VIN
V+ = 10 V
CD4011
VOUT
R3
1 MΩ
1/3 L144
VLOW
R4
1 MΩ
Vout = "LOW" WHEN:
VHIGH > VIN > VLOW
DIRECT CMOS OUTPUT
PD = 290 μW
V− = −10V

Fig. 16-13

Circuit Notes

The detector uses three sections of an L144 and a DC4011 type CMOS NAND gate to make a very low power voltage monitor. If the input voltage, VIN, is above VHIGH or below VLOW, the output will be a logical high. If (and only if) the input is between the limits will the output be low. The 1 megohm resistors R1, R2, R3, and R4 translate the bipolar ±10V swing of the op amps to a 0 to 10V swing acceptable to the ground-referenced CMOS logic.

OPPOSITE POLARITY INPUT VOLTAGE COMPARATOR

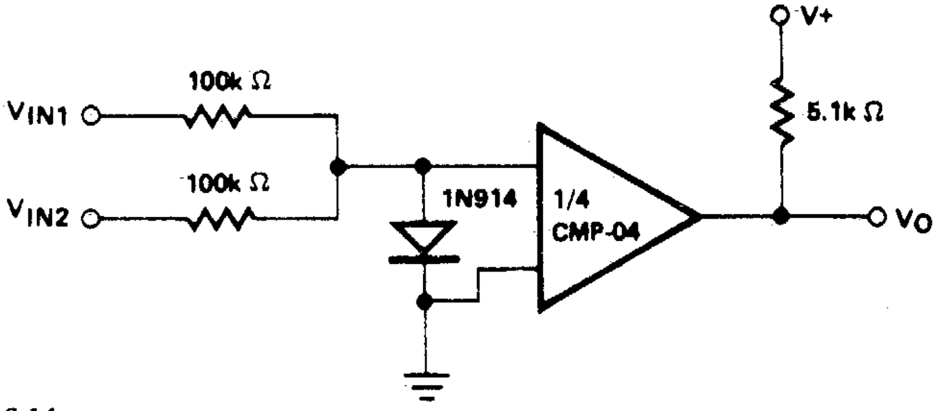

V+
100k Ω
VIN1
100k Ω
VIN2
5.1k Ω
1N914
1/4
CMP-04
VO

Fig. 16-14

LIMIT COMPARATOR

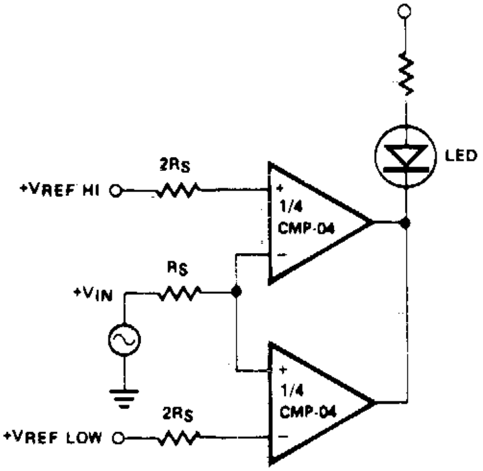

LED
2RS
+VREF HI
+
1/4
CMP-04
−
RS
+VIN
+
1/4
CMP-04
2RS
+VREF LOW
−

Fig. 16-15

COMPARATOR CLOCK CIRCUIT

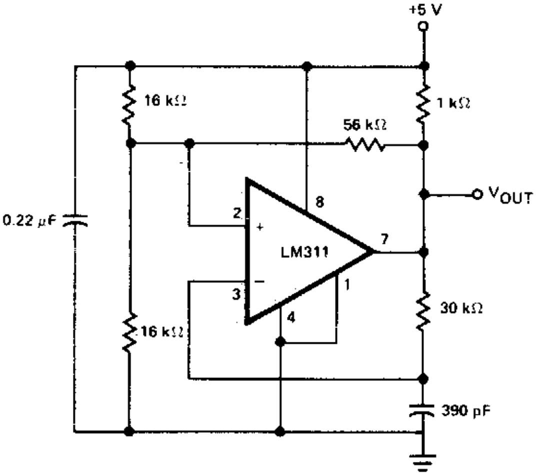

+5 V
16 kΩ
1 kΩ
56 kΩ
8
VOUT
0.22 µF
2
+
LM311
7
3
−
1
4
30 kΩ
16 kΩ
390 pF

Fig. 16-16

DOUBLE-ENDED LIMIT COMPARATOR

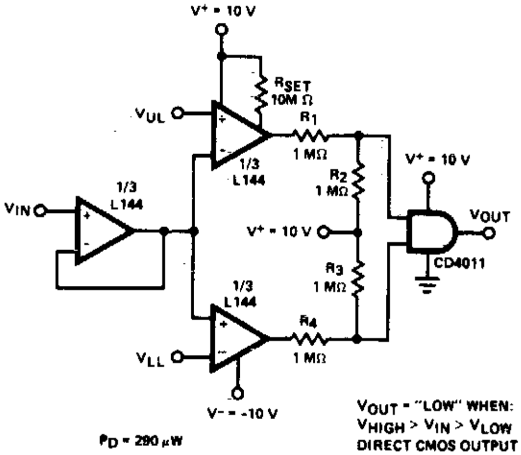

V+ = 10 V
RSET
10M Ω
R1
VUL
+
−
1/3
L144
1 MΩ
V+ = 10 V
R2
1 MΩ
1/3
L144
VIN
+
−
VOUT
V+ = 10 V
CD4011
R3
1 MΩ
1/3
L144
R4
+
−
VLL
1 MΩ
V− = −10 V
PD = 290 µW
VOUT = "LOW" WHEN:
VHIGH > VIN > VLOW
DIRECT CMOS OUTPUT

Fig. 16-17

LIMIT COMPARATOR

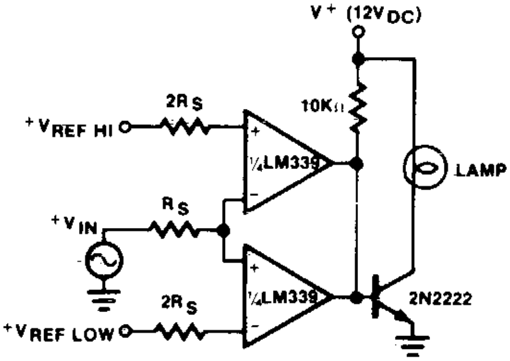
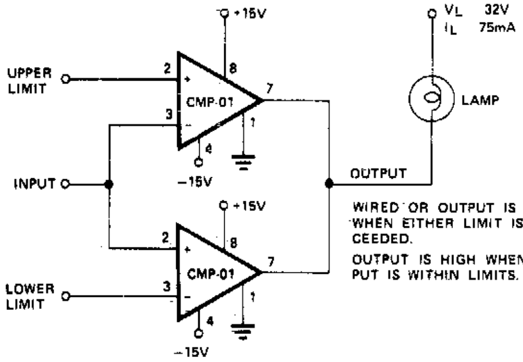

V+ (12VDC)
2RS
10KΩ
+VREF HI
+
¼LM339
−
LAMP
RS
+VIN
+
¼LM339
−
2N2222
2RS
+VREF LOW

Fig. 16-18

PRECISION, DUAL LIMIT, GO/NO GO TESTER

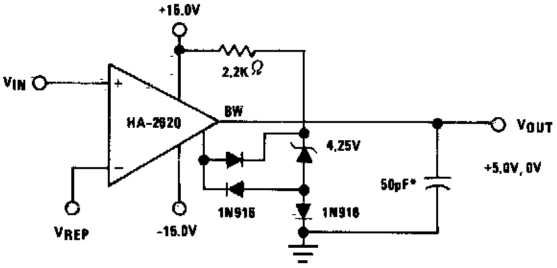

+15V
VL 32V
IL 75mA
UPPER LIMIT
2
8
CMP-01
7
3
1
4
LAMP
INPUT
−15V
OUTPUT
+15V
2
8
CMP-01
7
LOWER LIMIT
3
1
4
−15V

WIRED OR OUTPUT IS LOW WHEN EITHER LIMIT IS EXCEEDED.

OUTPUT IS HIGH WHEN INPUT IS WITHIN LIMITS.

Fig. 16-19

HIGH IMPEDANCE COMPARATOR

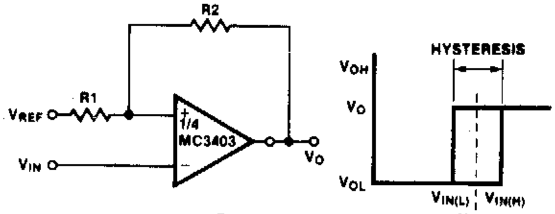

+15.0V
2.2K Ω
VIN
+
HA-2620
8W
VOUT
4.25V
+5.0V, 0V
50pF*
1N916
1N916
VREF
−15.0V

IC = HA-2620

Harris Semiconductor
Analog Data Book - 1984

Fig. 16-21

COMPARATOR WITH HYSTERESIS


R2
R1
VREF
VIN
1/4 MC3403
VO
HYSTERESIS
VOH
VO
VOL
VIN(L)
VIN(H)
VREF

$$V_{IN(L)} = \frac{R1}{R1 + R2}(V_{OL} - V_{REF}) + V_{REF}$$

$$V_{IN(H)} = \frac{R1}{R1 + R2}(V_{OH} - V_{REF}) + V_{REF}$$

$$H = \frac{R1}{R1 + R2}(V_{OH} - V_{OL})$$

Fig. 16-20

COMPARATOR

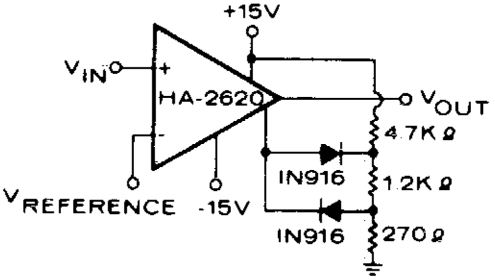

+15V
VIN
+
HA-2620
VOUT
4.7K Ω
IN916
1.2K Ω
VREFERENCE
-15V
IN916
270 Ω

Fig. 16-22

Circuit Notes

An operational amplifier is used as a comparator which is capable of driving approximately 10 logic gates.

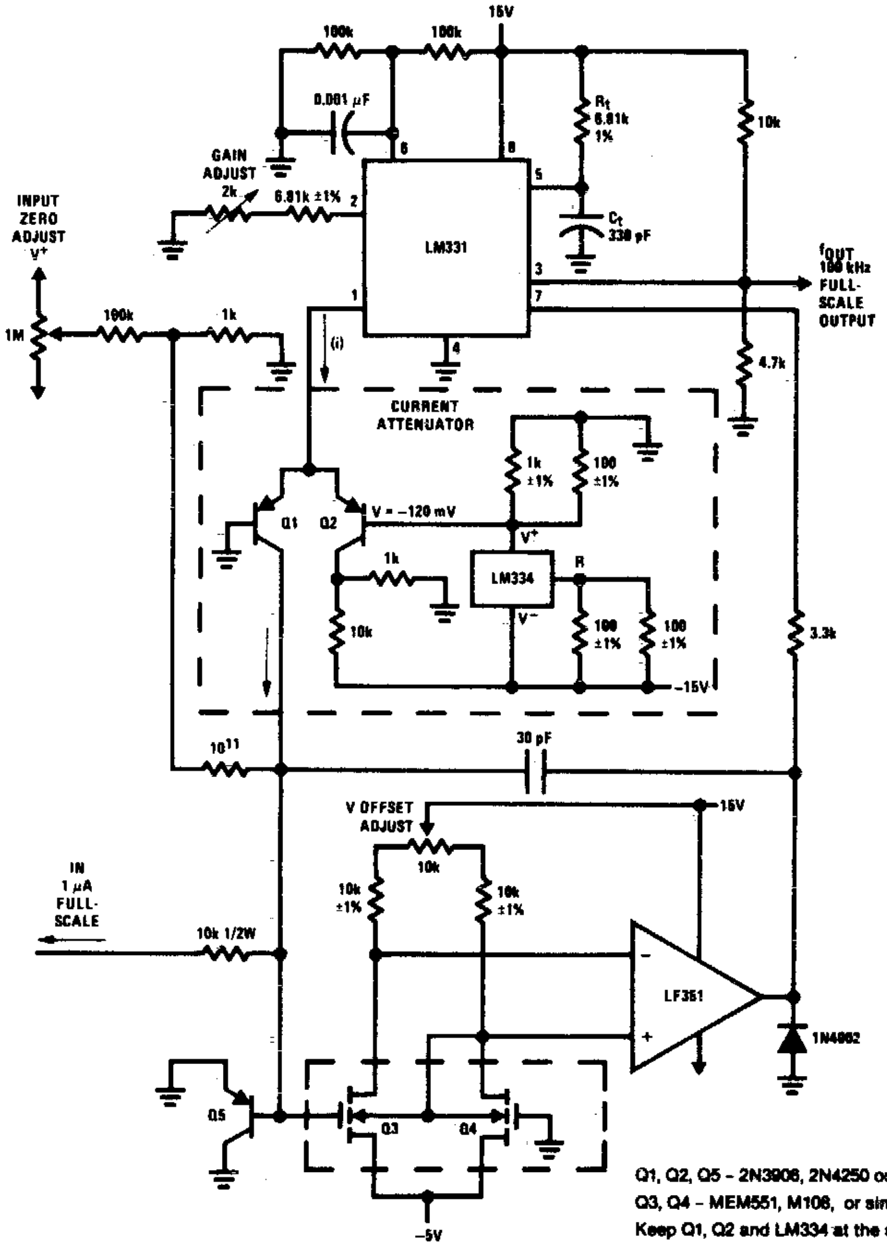
17

Converters

The sources of the following circuits are contained in the Sources section beginning on page 730. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

Picoampere-to-Frequency Converter
BCD-to-Analog Converter
Resistance-to-Voltage Converter
Low Cost, μP Interfaced, Temperature-to-Digital Converter
Hi-Lo Resistance-to-Voltage Converter
Current-to-Voltage Converter
Calculator-to-Stopwatch Converter
Power Voltage-to-Current Converter
High Impedance Precision Rectifier for Ac/Dc Converter
Wide Range Current-to-Frequency Converter
Ac-to-Dc Converter
Current-to-Voltage Converter with 1% Accuracy
Polarity Converter
Voltage-to-Current Converter
Wideband, High-Crest Factor, RMS-to-Dc Converter
Light Intensity-to-Frequency Converter
Ohms-to-Volts Converter
Temperature-to-Frequency Converter
Multiplexed BCD-to-Parallel BCD Converter
Fast Logarithmic Converter
Sine Wave-to-Square Wave Converter
Self Oscillating Flyback Converter
TTL-to-MOS Logic Converter
Picoampere-to-Voltage Converter with Gain

PICOAMPERE-TO-FREQUENCY CONVERTERS


15V
100k
100k
0.001 μF
6
8
R$_t$
6.81k
1%
10k
GAIN
ADJUST
2k
6.81k ±1%
2
5
C$_t$
330 pF
INPUT
ZERO
ADJUST
V$^+$
LM331
3
7
1
f$_{OUT}$
100 kHz
FULL-
SCALE
OUTPUT
1M
100k
1k
(i)
4
4.7k
CURRENT
ATTENUATOR
1k
±1%
100
±1%
Q1
Q2
V = −120 mV
V$^+$
1k
LM334
R
V$^-$
10k
100
±1%
100
±1%
−15V
3.3k
10^{11}
30 pF
V OFFSET
ADJUST
15V
10k
IN
1 μA
FULL-
SCALE
10k
±1%
10k
±1%
10k 1/2W
LF351
1N4002
Q5
Q3
Q4
−5V

Q1, Q2, Q5 – 2N3906, 2N4250 or similar
Q3, Q4 – MEM551, M108, or similar. See text
Keep Q1, Q2 and LM334 at the same temperature

Fig. 17-1

BCD-TO-ANALOG CONVERTER

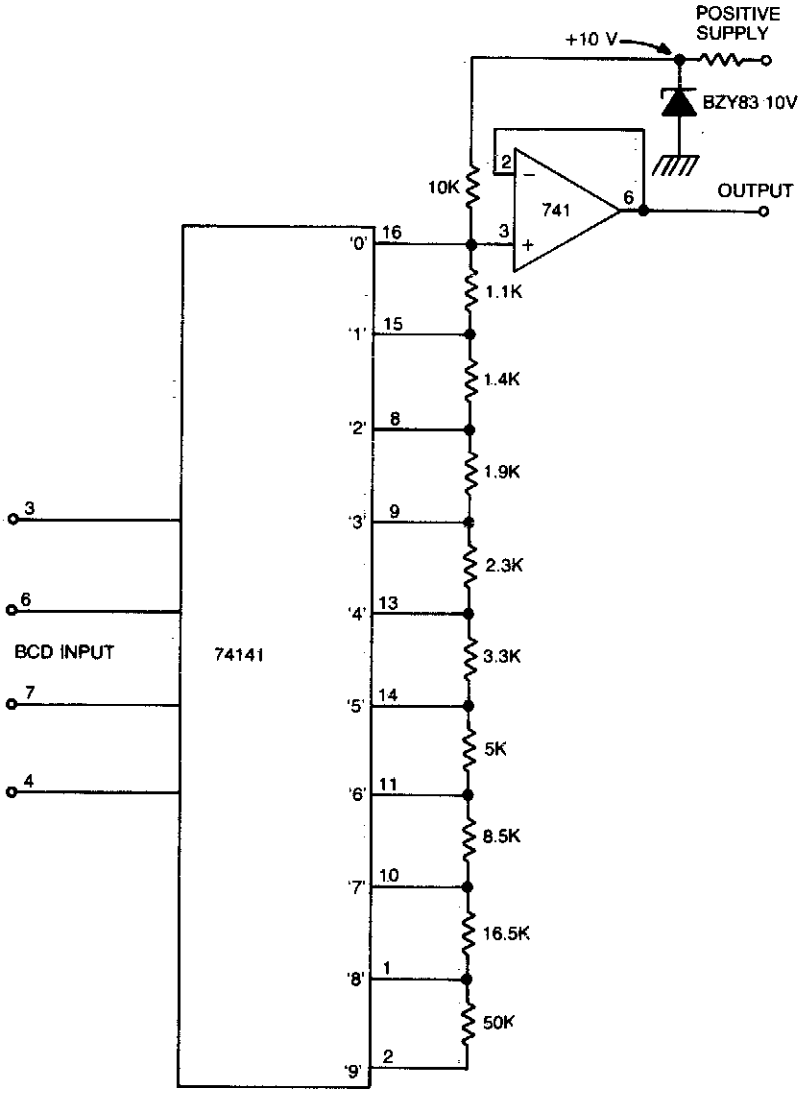
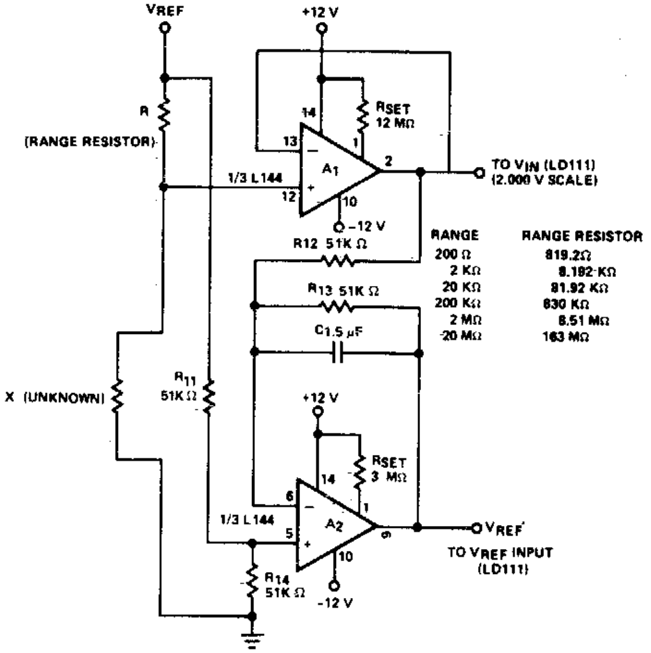

POSITIVE SUPPLY
+10 V
BZY83 10V
10K
741
2
3
6
OUTPUT
74141
'0'
16
'1'
15
'2'
8
'3'
9
'4'
13
'5'
14
'6'
11
'7'
10
'8'
1
'9'
2
1.1K
1.4K
1.9K
2.3K
3.3K
5K
8.5K
16.5K
50K
3
6
BCD INPUT
7
4

Fig. 17-2

Circuit Notes

This circuit will convert four-bit BCD into a variable voltage from 0-9 V in 1 V steps. The SN74141 is a Nixie driver, and has ten open-collector outputs. These are used to ground a selected point in the divider chain determined by the BCD code at the input, and so produce a corresponding voltage at the output. Accuracy of the circuit depends on the tolerance of the resistors and the accuracy of the reference voltage. However, presets can be used in the divider chain, with correct calibration. The 741 is used as a buffer.

RESISTANCE-TO-VOLTAGE CONVERTER


VREF
+12 V
R
(RANGE RESISTOR)
RSET 12 MΩ
14
13
1
2
A1
1/3 L144
12
10
-12 V
TO VIN (LD111) (2.000 V SCALE)
R12 51K Ω
R13 51K Ω
C1.5 μF
X (UNKNOWN)
R11 51K Ω
+12 V
RSET 3 MΩ
14
6
1
1/3 L144
5
A2
9
10
VREF'
TO VREF INPUT (LD111)
R14 51K Ω
-12 V

RANGE	RANGE RESISTOR
200 Ω	819.2Ω
2 KΩ	8.192 KΩ
20 KΩ	81.92 KΩ
200 KΩ	830 KΩ
2 MΩ	8.51 MΩ
20 MΩ	163 MΩ

Fig. 17-3

Circuit Notes

Circuit will measure accurately to 20 M when associated with a buffer amplifier (A1) having a low input bias current (I_{IN}) < 30 nA). The circuit uses two of the three amplifiers contained in the Siliconix L144 micropower triple op amp.

LOW-COST, μP INTERFACED, TEMPERATURE-TO-DIGITAL CONVERTER

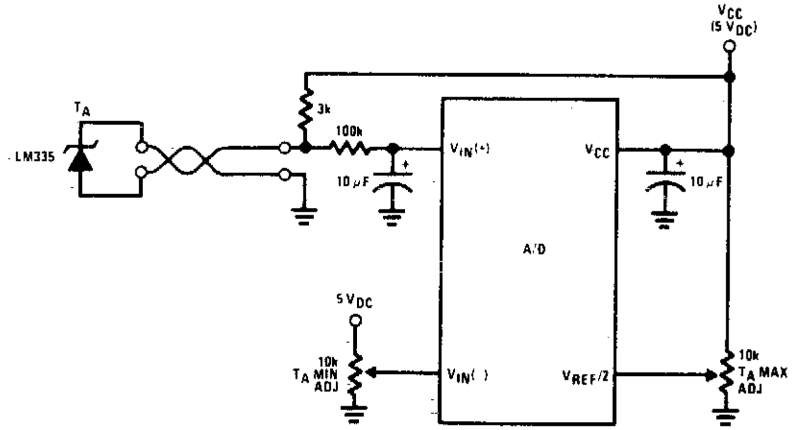
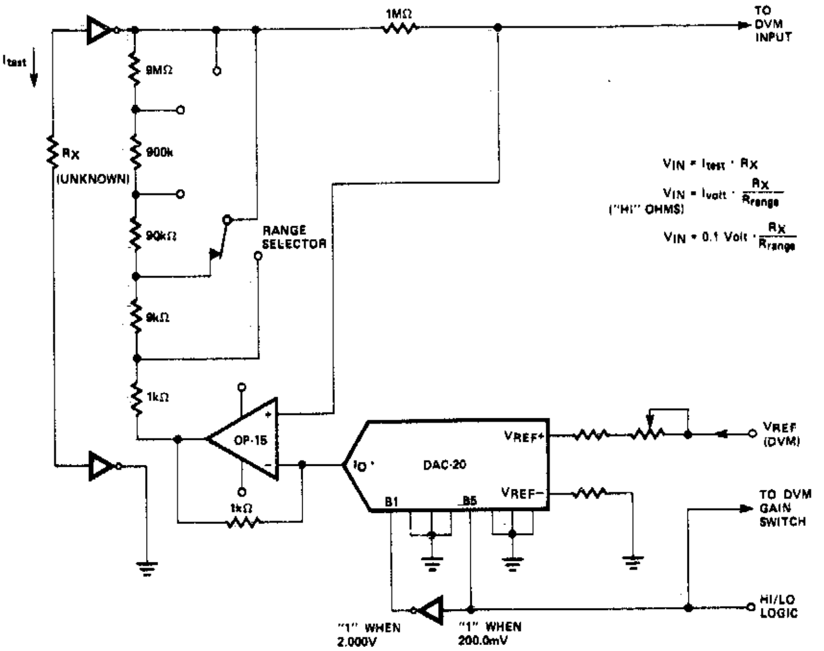

VCC (5 VDC)
TA
LM335
3k
100k
10 μF
VIN(+)
VCC
10 μF
A/D
5 VDC
10k
TA MIN ADJ
VIN(-)
VREF/2
10k
TA MAX ADJ

Fig. 17-4

HI-LO RESISTANCE-TO-VOLTAGE CONVERTER


1MΩ
TO DVM INPUT
Itest
9MΩ
900k
90kΩ
9kΩ
1kΩ
RX (UNKNOWN)
RANGE SELECTOR
OP-15
1kΩ
DAC-20
VREF+
VREF−
IO
B1
B5
VREF (DVM)
TO DVM GAIN SWITCH
HI/LO LOGIC
"1" WHEN 2.000V
"1" WHEN 200.0mV

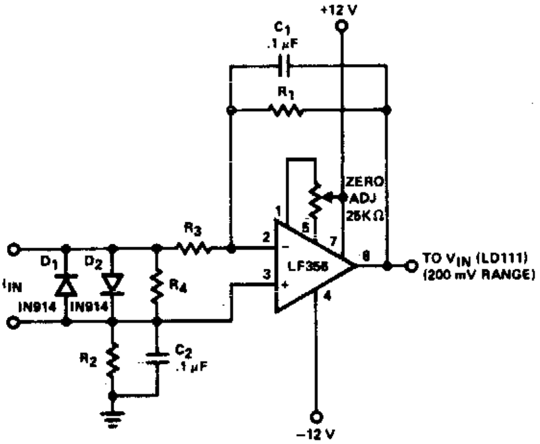
$V_{IN} = I_{test} \cdot R_X$

$V_{IN} = I_{volt} \cdot \frac{R_X}{R_{range}}$
("HI" OHMS)

$V_{IN} = 0.1 \text{ Volt} \cdot \frac{R_X}{R_{range}}$

Fig. 17-5

CURRENT-TO-VOLTAGE CONVERTER


+12 V
C1 .1 µF
R1
ZERO ADJ 25KΩ
R3
D1
D2
IIN
IN914
IN914
R4
R2
C2 .1µF
LF356
1
2
3
4
5
6
7
TO VIN (LD111) (200 mV RANGE)
−12 V

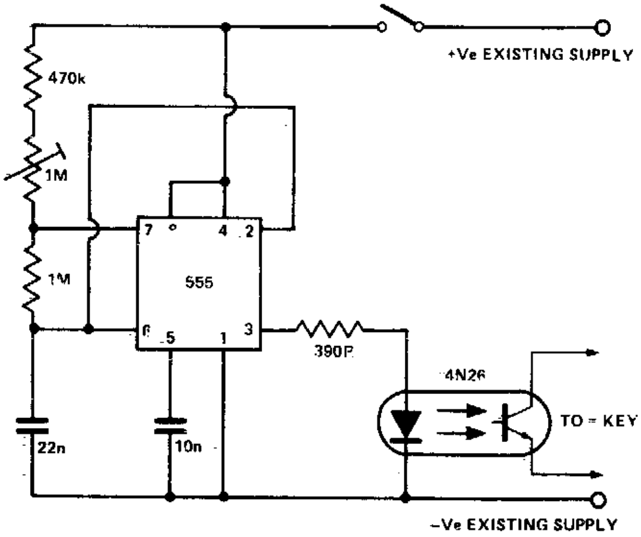
CURRENT RANGE	R_1	R_2	R_3	R_4
200 nA	500 KΩ	500 KΩ	0	∞
2 µA	50 KΩ	50 KΩ	0	∞
20 µA	5 KΩ	5 KΩ	0	∞
200 µA	1 KΩ	0	0	∞
2 mA	50 KΩ	0	5.0 K	10.0 Ω
20 mA	50 KΩ	0	5.0 K	1.0 Ω
200 mA	50 KΩ	0	5.0 K	.1 Ω
2 A	50 KΩ	0	5.0 K	.01 Ω

Circuit Notes

Converter features eight decades of current range. The circuit is intended to be used with the 200.0 mV range of a DVM.

Fig. 17-6

CALCULATOR-TO-STOPWATCH CONVERTER



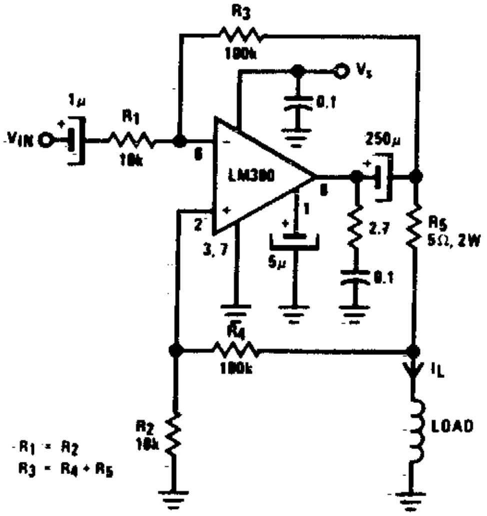
+Ve EXISTING SUPPLY
470k
1M
1M
7
4
2
555
6
5
1
3
390R
4N26
TO = KEY
22n
10n
–Ve EXISTING SUPPLY

Fig. 17-7

Circuit Notes

This circuit can be fitted to any calculator with an automatic constant to enable it to be used as a stop-watch. The 555 timer is set to run at a suitable frequency and connected to the existing calculator battery via the push-on push-off switch and the existing calculator on-off switch.

POWER VOLTAGE-TO-CURRENT CONVERTER



R3
100k
Vs
0.1
1μ
R1
10k
VIN
6
LM380
250μ
8
1
2
3, 7
5μ
2.7
0.1
R5
5Ω, 2W
R4
100k
IL
LOAD
R2
10k
R1 = R2
R3 = R4 + R5

Circuit Notes

Low cost converter is capable of supplying constant ac currents up to 1 A over variable loads.

Fig. 17-8

HIGH IMPEDANCE PRECISION RECTIFIER FOR AC/DC CONVERTER

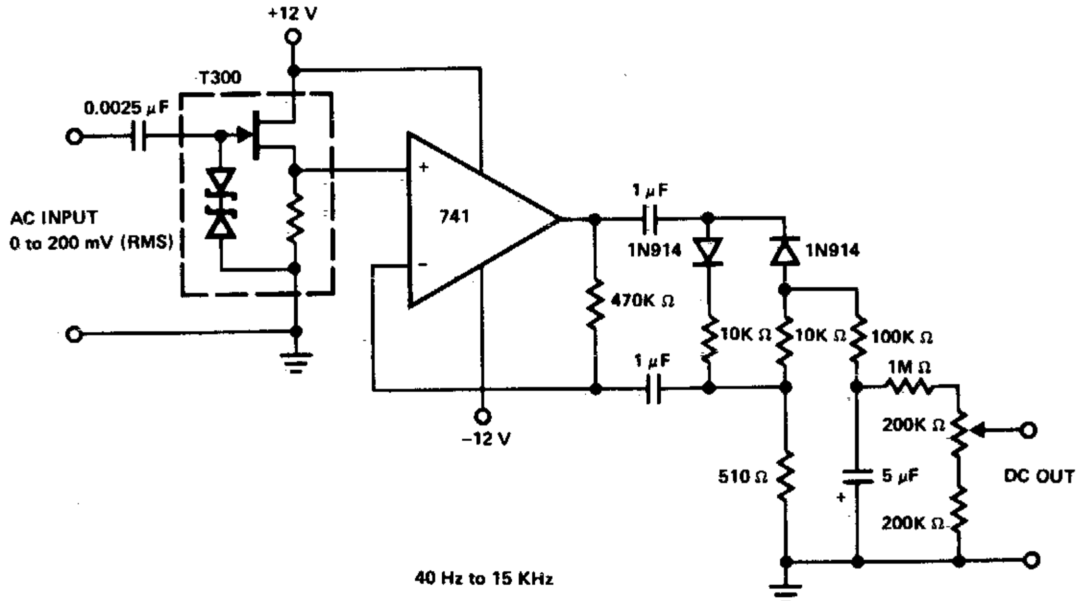
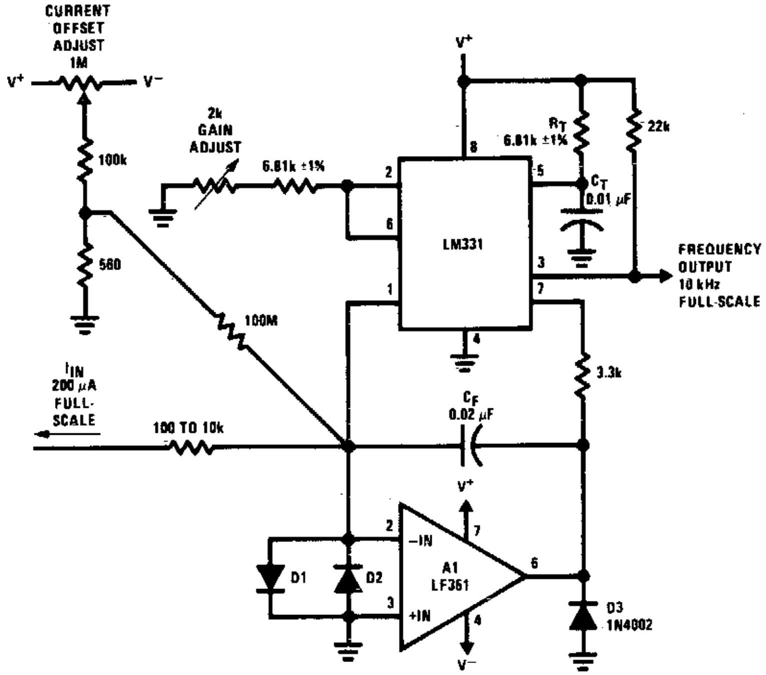

+12 V
T300
0.0025 μF
AC INPUT
0 to 200 mV (RMS)
+
741
−
1 μF
1N914
1N914
470K Ω
10K Ω
10K Ω
100K Ω
1 μF
1M Ω
200K Ω
−12 V
510 Ω
5 μF
+
DC OUT
200K Ω
40 Hz to 15 KHz

Fig. 17-9

WIDE-RANGE CURRENT-TO-FREQUENCY CONVERTER


CURRENT
OFFSET
ADJUST
1M
V+
V−
100k
2k
GAIN
ADJUST
6.81k ±1%
560
100M
V+
R_T
6.81k ±1%
22k
8
2
5
6
C_T
0.01 μF
LM331
3
FREQUENCY
OUTPUT
10 kHz
FULL-SCALE
1
7
4
3.3k
I_IN
200 μA
FULL-
SCALE
100 TO 10k
C_F
0.02 μF
V+
7
2
−IN
A1
LF361
6
3
+IN
4
V−
D1
D2
D3
1N4002

D1, D2 = 1N457, 1N484, or similar low-leakage planar diode

Fig. 17-10

AC-TO-DC CONVERTER

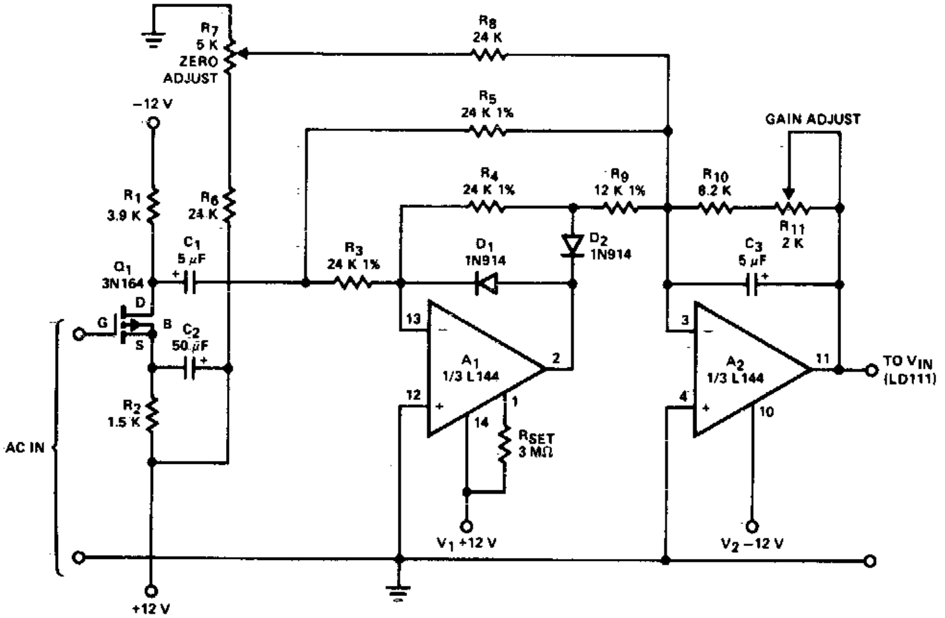
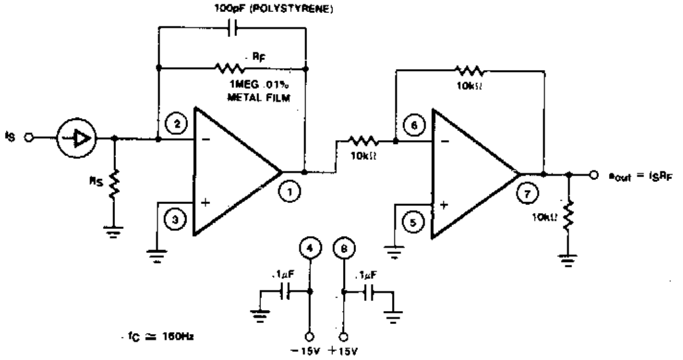

R7
5 K
ZERO
ADJUST
R8
24 K
R5
24 K 1%
GAIN ADJUST
−12 V
R1
3.9 K
R6
24 K
R4
24 K 1%
R9
12 K 1%
R10
8.2 K
R11
2 K
C1
5 μF
Q1
3N164
D
G
B
S
C2
50 μF
R3
24 K 1%
D1
1N914
D2
1N914
C3
5 μF
13
A1
1/3 L144
2
12
14
1
RSET
3 MΩ
3
A2
1/3 L144
11
4
10
TO VIN
(LD111)
R2
1.5 K
AC IN
V1 +12 V
V2 −12 V
+12 V

Fig. 17-11

Circuit Notes

This circuit includes a PMOS enhancement-mode FET input buffer amplifier, coupled to a classical absolute value circuit which essentially eliminates the effect of the forward voltage drop across diodes D1 and D2.

CURRENT-TO-VOLTAGE CONVERTER WITH 1% Accuracy


100pF (POLYSTYRENE)
RF
1MEG .01%
METAL FILM
10kΩ
IS
RS
2
3
1
10kΩ
6
5
7
eout = ISRF
10kΩ
4
8
.1μF
.1μF
fC ≅ 160Hz
−15V +15V

IC = NE/SE55/1
(Sensitivity: 1 Volt per Microamp)

Fig. 17-12

Circuit Notes

A filter removes the dc component of the rectified ac, which is then scaled to RMS. The output is linear from 40 Hz to 10 kHz or higher.

POLARITY CONVERTER

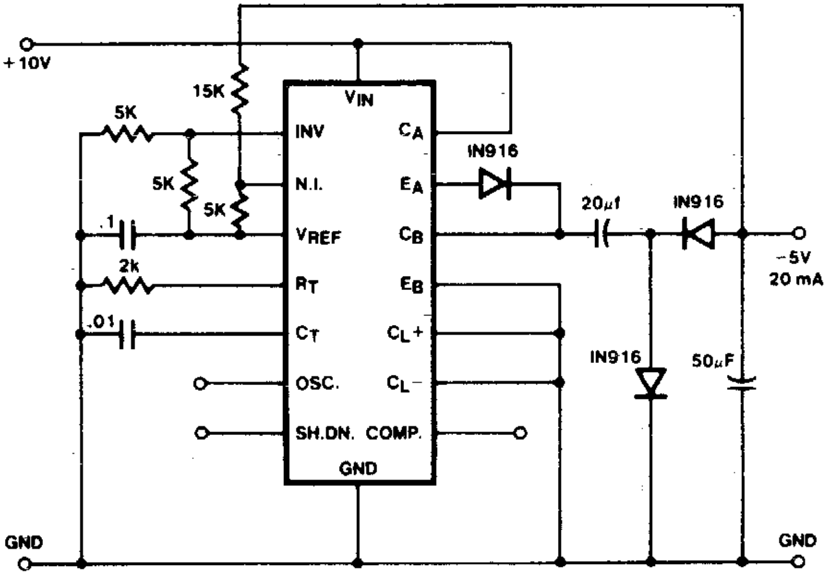
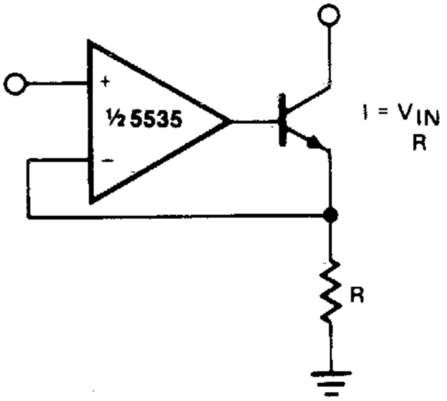

+10V
15K
5K
5K
5K
.1
2k
.01
VIN
INV
N.I.
VREF
RT
CT
OSC.
SH.DN.
COMP.
GND
CA
EA
CB
EB
CL+
CL−
IN916
20μf
IN916
−5V
20 mA
IN916
50μF
GND
GND

Fig. 17-13

Circuit Notes

The capacitor-diode output circuit is used here as a polarity converter to generate a −5 volt supply from +15 volts. This circuit is useful for an output current of up to 20 mA with no additional boost transistors required. Since the output transistors are current limited, no additional protection is necessary. Also, the lack of an inductor allows the circuit to be stabilized with only the output capacitor.

VOLTAGE-TO-CURRENT CONVERTER

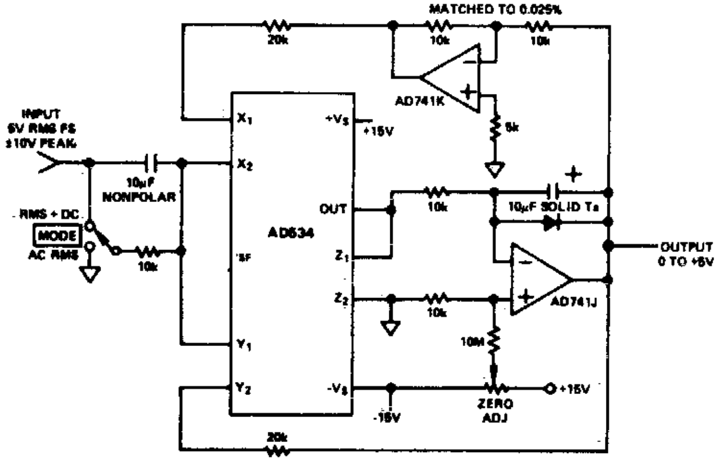

½5535
+
−
I = VIN / R
R

Circuit Notes

The current out is $I_{OUT} \cong V_{IN}/R$. For negative currents, a PNP can be used and, for better accuracy, a Darlington pair can be substituted for the transistor. With careful design, this circuit can be used to control currents of many amps. Unity gain compensation is necessary.

Fig. 17-14

WIDEBAND, HIGH-CREST FACTOR, RMS-TO-DC CONVERTER


MATCHED TO 0.025%
20k
10k
10k
AD741K
5k
INPUT
5V RMS FS
±10V PEAK
10µF
NONPOLAR
RMS + DC
MODE
AC RMS
10k
X_1
X_2
$+V_S$
+15V
OUT
AD534
SF
Z_1
Z_2
Y_1
Y_2
$-V_S$
-15V
10k
10µF SOLID Ta
OUTPUT
0 TO +5V
AD741J
10k
10M
+15V
ZERO
ADJ
20k

CALIBRATION PROCEDURE:

WITH 'MODE' SWITCH IN 'RMS + DC' POSITION, APPLY AN INPUT OF +1.00VDC. ADJUST ZERO UNTIL OUTPUT READS SAME AS INPUT. CHECK FOR INPUTS OF ±10V; OUTPUT SHOULD BE WITHIN ±0.05% (5mV).

ACCURACY IS MAINTAINED FROM 60Hz to 100kHz, AND IS TYPICALLY HIGH BY 0.5% AT 1MHz FOR V_{IN} = 4V RMS (SINE, SQUARE OR TRIANGULAR WAVE).

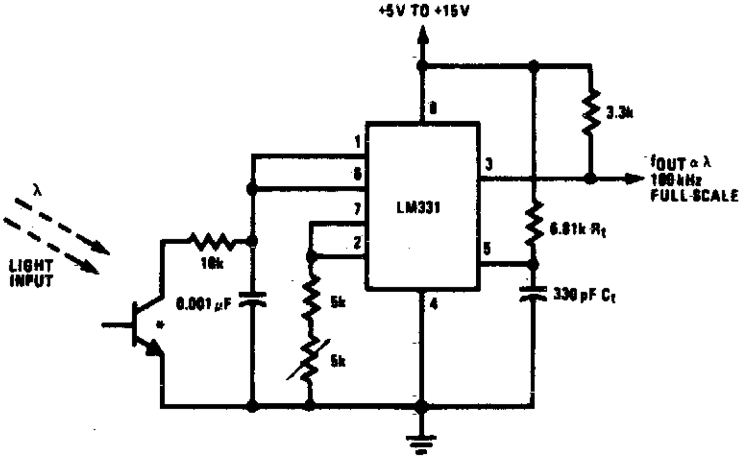
PROVIDED THAT THE PEAK INPUT IS NOT EXCEEDED, CREST-FACTORS UP TO AT LEAST TEN HAVE NO APPRECIABLE EFFECT ON ACCURACY.

INPUT IMPEDANCE IS ABOUT 10kΩ; FOR HIGH (10MΩ) IMPEDANCE, REMOVE MODE SWITCH AND INPUT COUPLING COMPONENTS.

FOR GUARANTEED SPECIFICATIONS THE AD536A AND AD636 IS OFFERED AS A SINGLE PACKAGE RMS-TO-DC CONVERTER.

Fig. 17-15

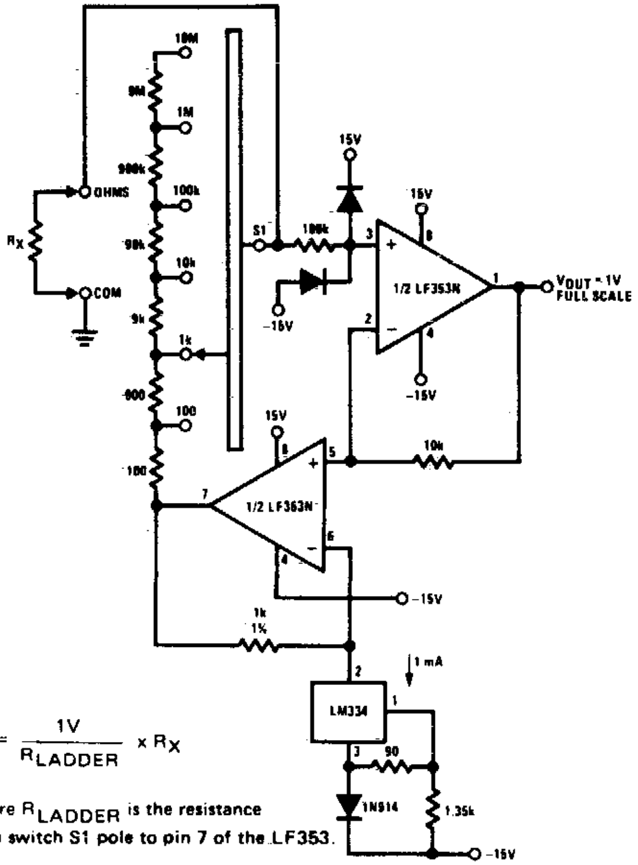
LIGHT INTENSITY-TO-FREQUENCY CONVERTER


+5V TO +15V
8
3.3k
1
6
7
2
3
5
4
LM331
$f_{OUT} \propto \lambda$
100 kHz
FULL-SCALE
λ
LIGHT
INPUT
10k
0.001 µF
*
5k
5k
6.81k R_t
330 pF C_t

*L14F-1, L14G-1 or L14H-1, photo transistor (General Electric Co.) or similar

Fig. 17-16

OHMS-TO-VOLTS CONVERTER


10M
9M
1M
900k
100k
OHMS
R_X
COM
90k
10k
9k
1k
900
100
100
S1
100k
15V
15V
3
8
+
1/2 LF353N
1
V_{OUT} = 1V FULL SCALE
−15V
2
−
4
−15V
15V
8
5
+
10k
7
1/2 LF353N
6
−
4
−15V
1k 1%
1 mA
2
LM334
1
3
90
1N914
1.35k
−15V

$$V_O = \frac{1V}{R_{LADDER}} \times R_X$$

Where R_{LADDER} is the resistance from switch S1 pole to pin 7 of the LF353.

Fig. 17-17

TEMPERATURE-TO-FREQUENCY CONVERTER

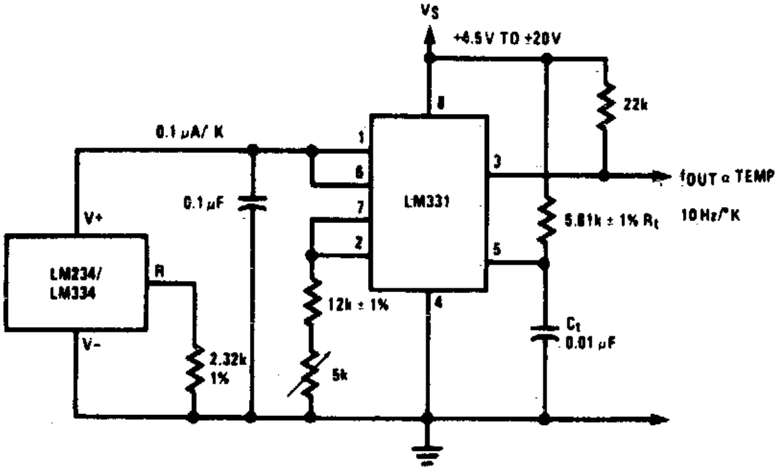
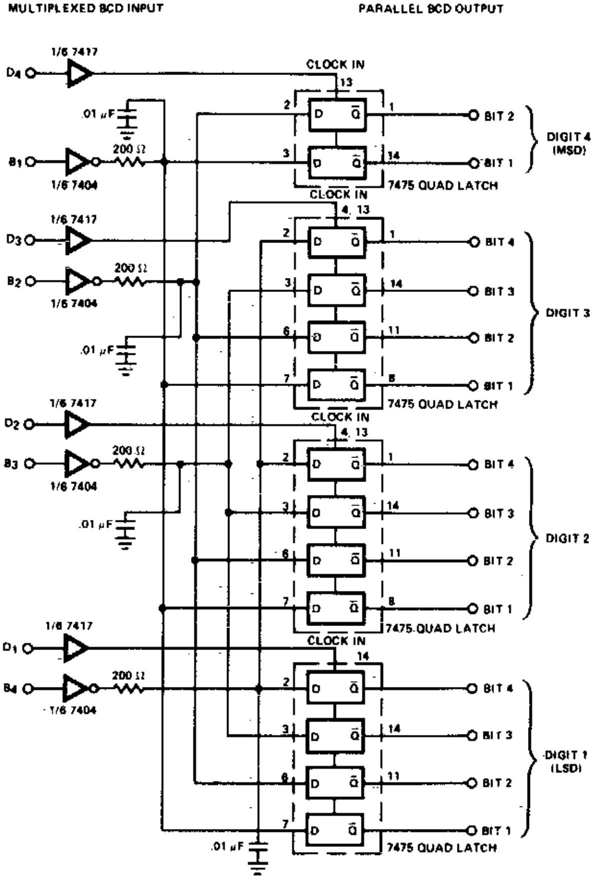

V_S
+4.5V TO ±20V
8
22k
0.1 μA/ K
1
6
3
f_{OUT} α TEMP
0.1 μF
7
LM331
5.81k ± 1% R_t
10 Hz/°K
V+
2
5
LM234/ LM334
R
12k ± 1%
4
C_t 0.01 μF
V-
2.32k 1%
5k

Fig. 17-18

MULTIPLEXED BCD-TO-PARALLEL BCD CONVERTER



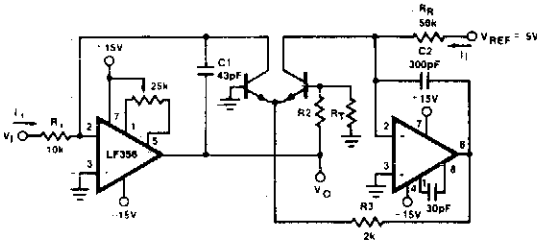
MULTIPLEXED BCD INPUT
PARALLEL BCD OUTPUT
1/6 7417
D4
.01 μF
200 Ω
B1
1/6 7404
CLOCK IN
13
D
Q̄
BIT 2
BIT 1
DIGIT 4 (MSD)
7475 QUAD LATCH
4, 13
1/6 7417
D3
200 Ω
B2
1/6 7404
.01 μF
BIT 4
BIT 3
BIT 2
BIT 1
DIGIT 3
7475 QUAD LATCH
1/6 7417
D2
200 Ω
B3
1/6 7404
.01 μF
DIGIT 2
7475 QUAD LATCH
1/6 7417
D1
200 Ω
B4
1/6 7404
14
DIGIT 1 (LSD)
.01 μF
7475 QUAD LATCH

Circuit Notes

Converter consists of four quad bistable latches activated in the proper sequence by the digit strobe output of the LD110. The complemented outputs (Q) of the quad latch set reflects the state of the bit outputs when the digit strobe goes high. It will maintain this state when the digit strobe goes low.

Fig. 17-19

FAST LOGARITHMIC CONVERTER



RR
50k
VREF = 5V
C2
300pF
IH
+15V
C1
43pF
25k
II
RI
VI
10k
LF356
-15V
R2
RT
+15V
VO
R3
2k
-15V
30pF

$$*|V_{OUT}| = \left[1 + \frac{R2}{R}\right]\frac{kT}{q}\ln V_I\left[\frac{R_r}{V_{REF}\,R_I}\right] = \log V_I\,\frac{1}{R_I\,I_H}$$

*R2 = 15.7 k, R_T = 1k, 0.3%/°C (for temperature compensation)

- Dynamic range: 100μA ≤ I_I ≤ 1mA (5 decades, $|V_O|$ = 1V/decades)
- Transient response: 3μs for ΔI_I = decades
- C1, C2, R2, R3: added dynamic compensation
- V_{OS} adjust the LF356 to minimize quiescent error
- R_T: Tel Labs type Q81 + 0.3%/°C.

Fig. 17-20

SINE WAVE-TO-SQUARE WAVE CONVERTER

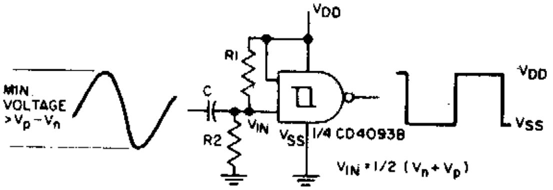

VDD
R1
MIN VOLTAGE >Vp - Vn
C
R2
VIN
VSS
1/4 CD4093B
VDD
VSS
VIN = 1/2 (Vn + Vp)

Fig. 17-21

Circuit Notes

The sine input is ac coupled by capacitor C; R1 and R2 bias the input midway between V_n and V_p, the input threshold voltages, to provide a square wave at the output.

SELF OSCILLATING FLYBACK CONVERTER

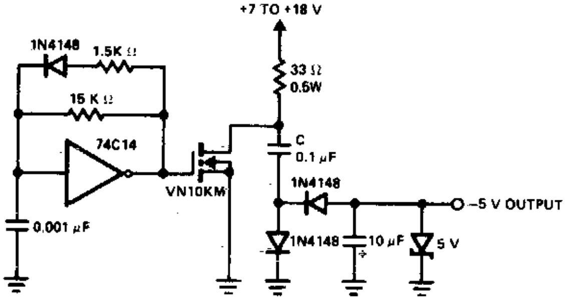

+7 TO +18 V
1N4148
1.5K Ω
15 K Ω
33 Ω
0.5W
74C14
C
0.1 μF
VN10KM
1N4148
–5 V OUTPUT
0.001 μF
1N4148
10 μF
5 V

Fig. 17-22

Circuit Notes

A low-power converter suitable for deriving a higher voltage from a main system rail in an on-board application. With the transformer shown, the operating frequency is 250 kHz. Z1 serves as a dissipative voltage regulator for the output and also clips the drain voltage to a level below the rated VMOS breakdown voltage.

TTL-TO-MOS LOGIC CONVERTER

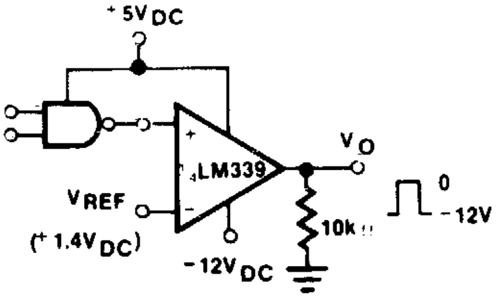

+5VDC
+
¼LM339
–
VO
VREF
(+1.4VDC)
-12VDC
10k Ω
0
-12V

Fig. 17-23

PICOAMPERE-TO-VOLTAGE CONVERTER WITH GAIN

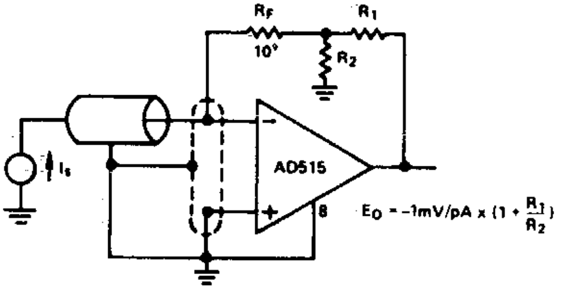

RF
10⁹
R1
R2
Is
–
+
AD515
8
EO = –1mV/pA x (1 + R1/R2)

Fig. 17-24

18

Crossover Networks

The sources of the following circuits are contained in the Sources section beginning on page 730. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

Active Crossover Network
Asymmetrical Third Order Butterworth Active Crossover Network
Third Order Butterworth Crossover Network

ACTIVE CROSSOVER NETWORK

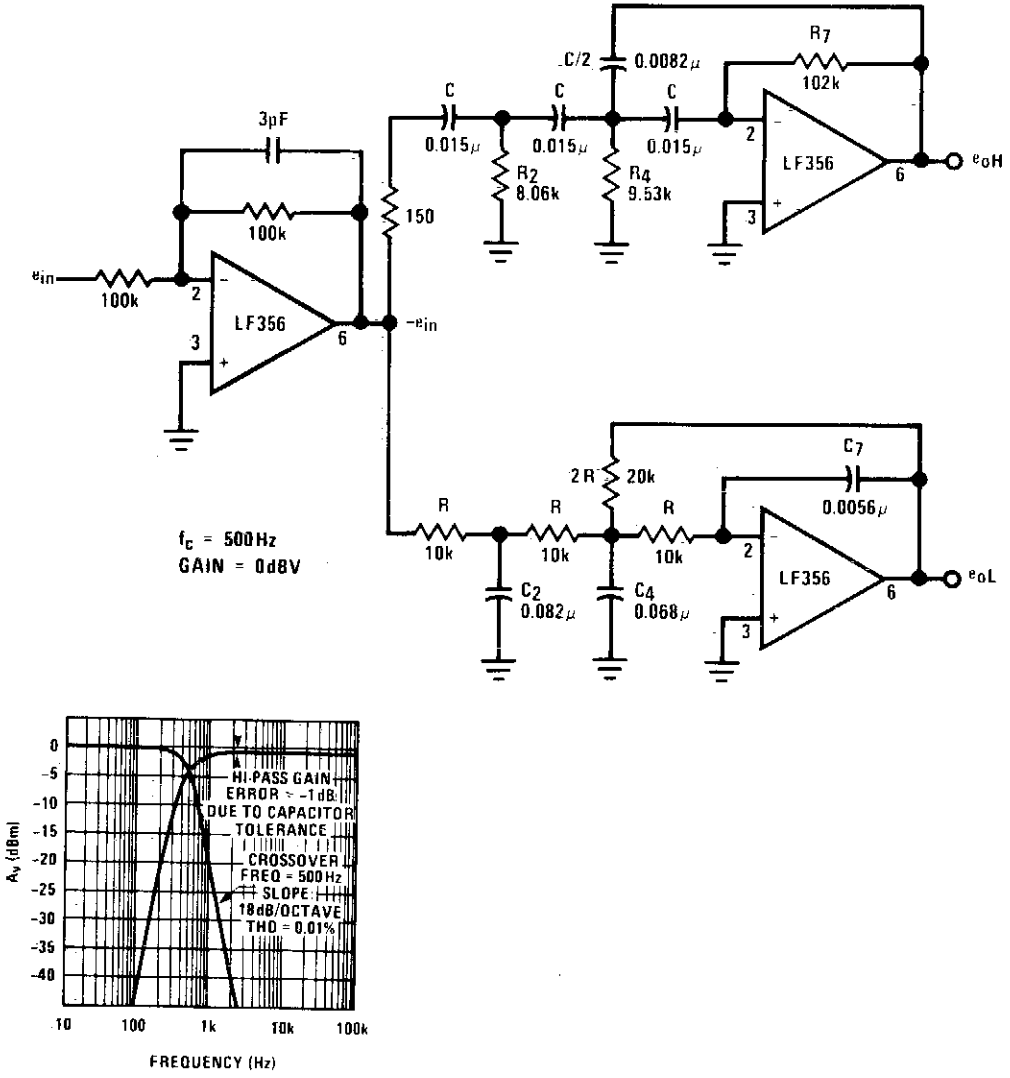
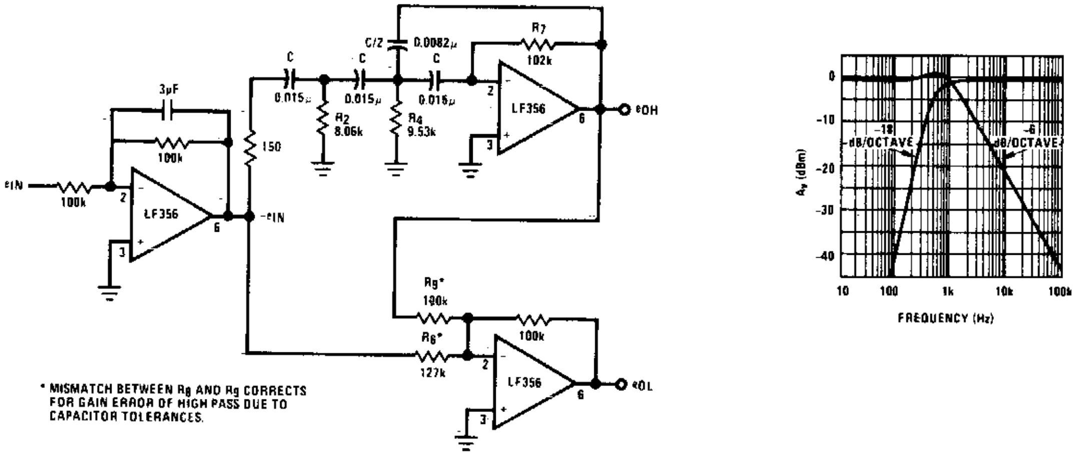

3pF
100k
ein
100k
2
3
LF356
6
-ein
150
C
0.015μ
C
0.015μ
C/2
0.0082μ
C
0.015μ
R2
8.06k
R4
9.53k
R7
102k
2
3
LF356
6
eoH
2R
20k
C7
0.0056μ
R
10k
R
10k
R
10k
C2
0.082μ
C4
0.068μ
2
3
LF356
6
eoL
fc = 500Hz
GAIN = 0dBV
0
-5
-10
-15
-20
-25
-30
-35
-40
Av (dBm)
10
100
1k
10k
100k
FREQUENCY (Hz)
HI-PASS GAIN
ERROR ≈ -1dB
DUE TO CAPACITOR
TOLERANCE
CROSSOVER
FREQ = 500Hz
SLOPE:
18dB/OCTAVE
THD = 0.01%

Fig. 18-1

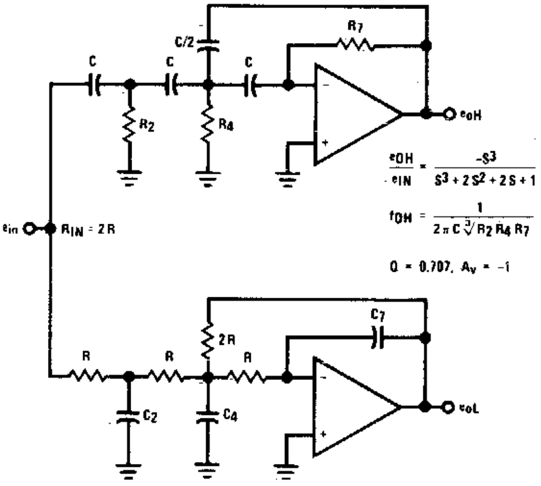
ASYMMETRICAL THIRD ORDER BUTTERWORTH ACTIVE CROSSOVER NETWORK


3μF
100k
eIN
100k
2
LF356
3
6
-eIN
150
C
0.015μ
R2
8.06k
C
0.015μ
C/2
0.0082μ
R4
9.53k
C
0.016μ
R7
102k
2
LF356
3
6
eOH
R9*
100k
R8*
127k
100k
2
LF356
3
6
eOL
0
-10
-20
-30
-40
Av (dBm)
-18 dB/OCTAVE
-6 dB/OCTAVE
10
100
1k
10k
100k
FREQUENCY (Hz)

* MISMATCH BETWEEN R8 AND R9 CORRECTS FOR GAIN ERROR OF HIGH PASS DUE TO CAPACITOR TOLERANCES.

Fig. 18-2

THIRD ORDER BUTTERWORTH CROSSOVER NETWORK


C
C/2
C
C
R7
R2
R4
eOH
ein
RIN = 2R
2R
C7
R
R
R
C2
C4
eOL

$$\frac{e_{OH}}{e_{IN}} = \frac{-S^3}{S^3+2S^2+2S+1}$$

$$f_{OH} = \frac{1}{2\pi C\sqrt[3]{R_2R_4R_7}}$$

$Q = 0.707,\ A_V = -1$

Fig. 18-3

$C_2 = \frac{2.4553}{2\pi f_{OL} R}$ $R_2 = \frac{0.4074}{2\pi f_{OH} C}$ $\frac{e_{OL}}{e_{IN}} = \frac{-1}{S^3+2S^2+2S+1}$

$C_4 = \frac{2.1089}{2\pi f_{OL} R}$ $R_4 = \frac{0.4742}{2\pi f_{OH} C}$ $f_{OL} = \frac{1}{2\pi R\sqrt[3]{C_2C_4C_7}}$

$C_7 = \frac{0.1931}{2\pi f_{OL} R}$ $R_7 = \frac{5.1766}{2\pi f_{OH} C}$ $Q = 0.707,\ A_V = -1$

19

Crystal Oscillators

The sources of the following circuits are contained in the Sources section beginning on page 730. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

High Frequency Crystal Oscillator
Overtone Crystal Oscillator
Overtone Crystal Oscillator
TTL Oscillator for 1 MHz-10 MHz
Crystal Checker
96 MHz Crystal Oscillator
Simple TTL Crystal Oscillator
Crystal Oscillator
Overtone Crystal Oscillator
Schmitt Trigger Crystal Oscillator
50 MHz-150 MHz Overtone Oscillator
Fifth Overtone Oscillator
Crystal Controlled Butler Oscillator
Overtone Oscillator with Crystal Switching
Crystal Oscillator
Crystal Oscillator/Doubler
Low Frequency Crystal Oscillator
Crystal Oscillator
100 kHz Crystal Calibrator
Third Overtone Crystal Oscillator
Crystal Checker
CMOS Crystal Oscillator
Temperature-Compensated Crystal Oscillator
Crystal Controlled Transistor Oscillator
Pierce Harmonic Oscillator
Colpitts Harmonic Oscillator
International Crystal OF-1 LO Oscillator
Butler Emitter Follower Oscillator
Colpitts Harmonic Oscillator
Butler Emitter Follower Oscillator
Butler Common Base Oscillator
Pierce Harmonic Oscillator
Tube Type Crystal Oscillator
Precision Clock Generator
Miller Oscillator
Butler Emitter Follower Oscillator
Colpitts Oscillator
Crystal-Controlled Oscillator
Pierce Oscillator
Butler Aperiodic Oscillator
Parallel-mode Aperiodic Crystal Oscillator
International Crystal OF-1 HI Oscillator
Standard Crystal Oscillator for 1 MHz
TTL-Compatible Crystal Oscillator
Crystal Controlled Sine Wave Oscillator
Crystal Oscillator
Stable Low Frequency Crystal Oscillator
JFET Pierce Crystal Oscillator
CMOS Oscillator
Pierce Harmonic Oscillator

HIGH FREQUENCY CRYSTAL OSCILLATOR

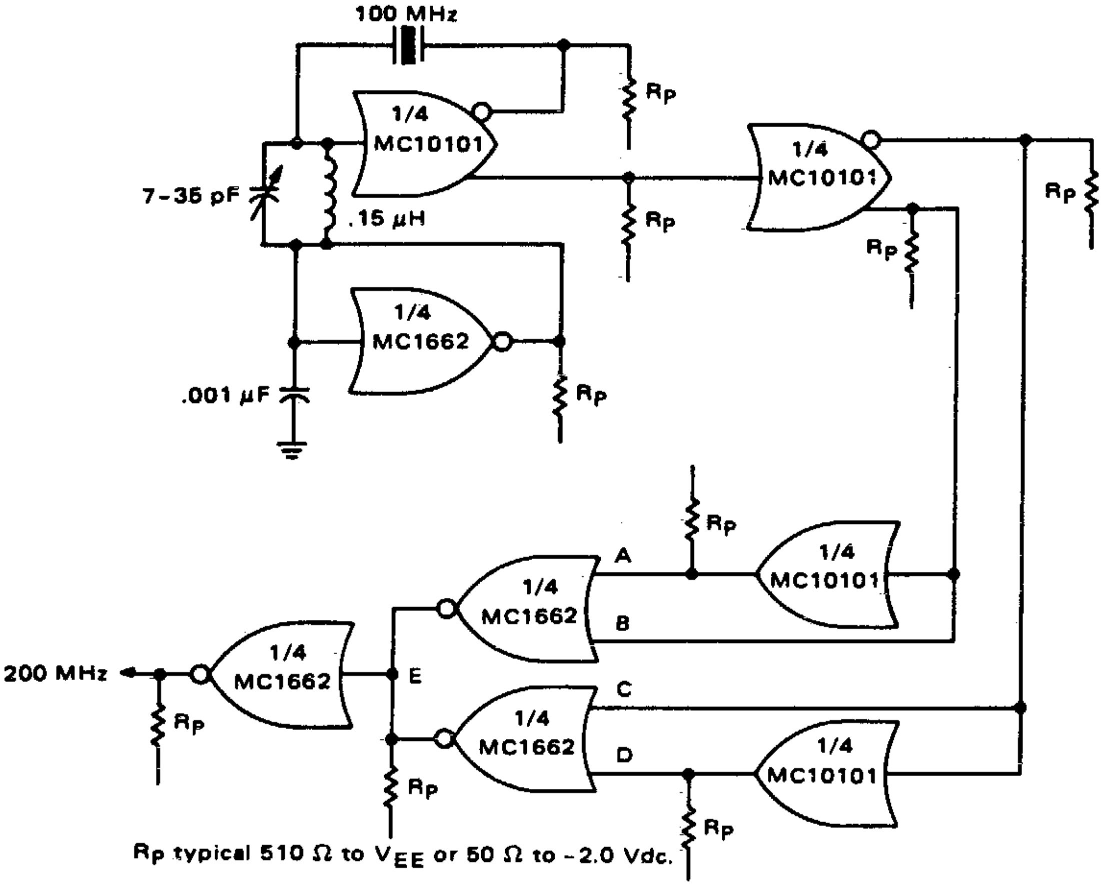
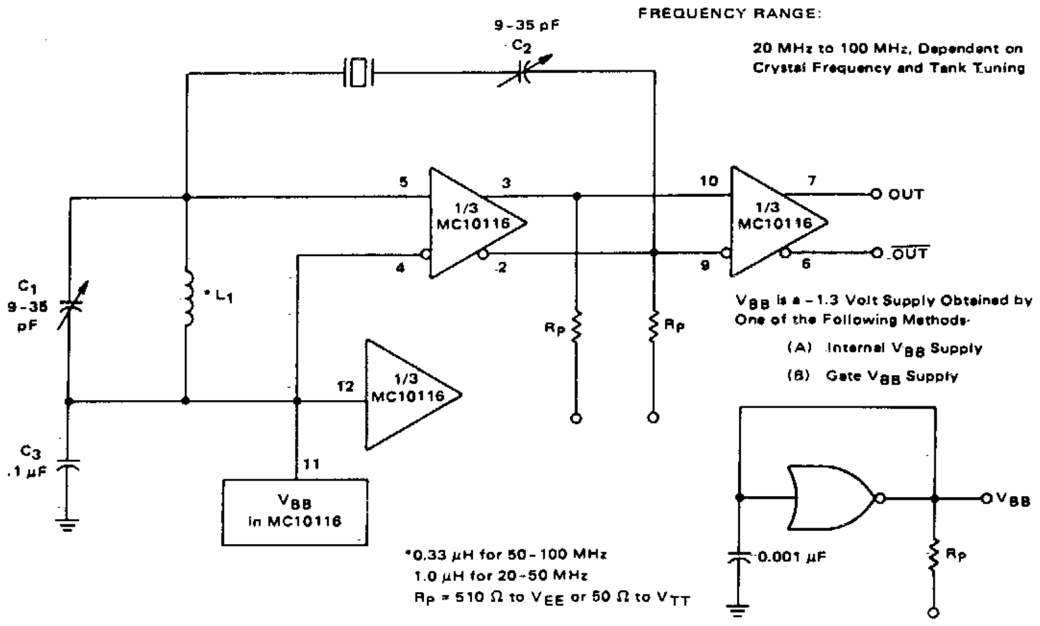

100 MHz
1/4 MC10101
Rp
7-35 pF
.15 μH
1/4 MC10101
Rp
Rp
Rp
1/4 MC1662
.001 μF
Rp
Rp
A
1/4 MC10101
1/4 MC1662
B
200 MHz
1/4 MC1662
E
Rp
C
1/4 MC1662
D
1/4 MC10101
Rp
Rp
Rp typical 510 Ω to V_EE or 50 Ω to -2.0 Vdc.

Fig. 19-1

Circuit Notes

One section of the MC10101 is connected as a 100 MHz crystal oscillator with the crystal in series with the feedback loop. The LC tank circuit tunes the 100 MHz harmonic of the crystal and may be used to calibrate the circuit to the exact frequency. A second section of the MC10101 buffers the crystal oscillator and gives complementary 100 MHz signals. The frequency doubler consists of two MC10101 gates as phase shifters and two MC1662 NOR gates. For a 50% duty cycle at the output, the delay to the true and complement 100 MHz signals should be 90°. This may be built precisely with 2.5 ns delay lines for the 200 MHz output or approximated by the two MC10101 gates as shown.

OVERTONE CRYSTAL OSCILLATOR


9-35 pF
C_2
5
3
1/3 MC10116
10
7
OUT
$\overline{OUT}$
4
2
9
6
C_1 9-35 pF
*L_1
R_P
R_P
12
1/3 MC10116
11
V_{BB} in MC10116
C_3 .1 µF
0.001 µF
V_{BB}
R_P

FREQUENCY RANGE:

20 MHz to 100 MHz, Dependent on Crystal Frequency and Tank Tuning

V_{BB} is a −1.3 Volt Supply Obtained by One of the Following Methods·

(A) Internal V_{BB} Supply

(B) Gate V_{BB} Supply

*0.33 µH for 50-100 MHz
1.0 µH for 20-50 MHz
R_P = 510 Ω to V_{EE} or 50 Ω to V_{TT}

Fig. 19-2

Circuit Notes

This circuit employs an adjustable resonant tank circuit which insures operation at the desired crystal overtone. C1 and L1 form the resonant tank circuit, which with the values specified as a resonant frequency adjustable from approximately 50 MHz to 100 MHz. Overtone operation is accomplished by adjusting the tank circuit frequency at or near the desired frequency. The tank circuit exhibits a low impedance shunt to off-frequency oscillations and a high impedance to the desired frequency, allowing feedback from the output. Operation in this manner guarantees that the oscillator will always start at the correct overtone.

OVERTONE CRYSTAL OSCILLATOR

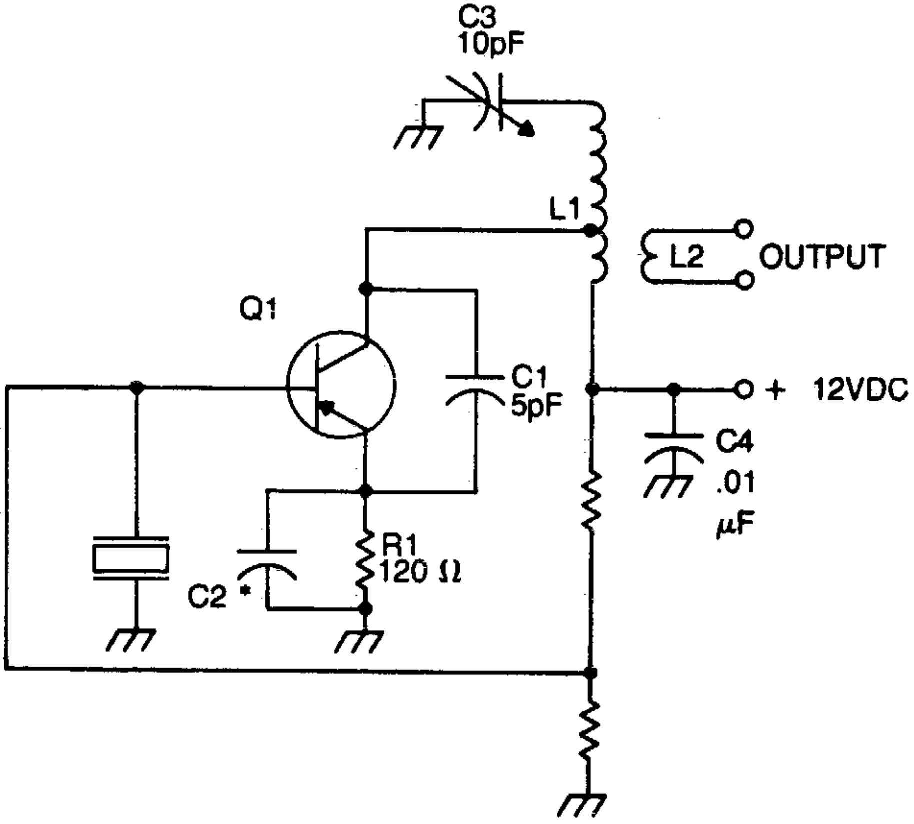


Fig. 19-3

Circuit Notes

The crystal element in this circuit is connected directly between the base and ground. Capacitor C1 is used to improve the feedback due to the internal capacitances of the transistor. This capacitor should be mounted as close as possible to the case of the transistor. The LC tank circuit in the collector of the transistor is tuned to the overtone frequency of the crystal. The emitter resistor capacitor must have a capacitive reactance of approximately 90 ohms at the frequency of operation. The tap on inductor L1 is used to match the impedance of the collector of the transistor. In most cases, the optimum placement of this tap is approximately one-third from the cold end of the coil. The placement of this tap is a trade-off between stability and maximum power output. The output signal is taken from a link coupling coil, L2, and operates by transformer action.

CRYSTAL CHECKER

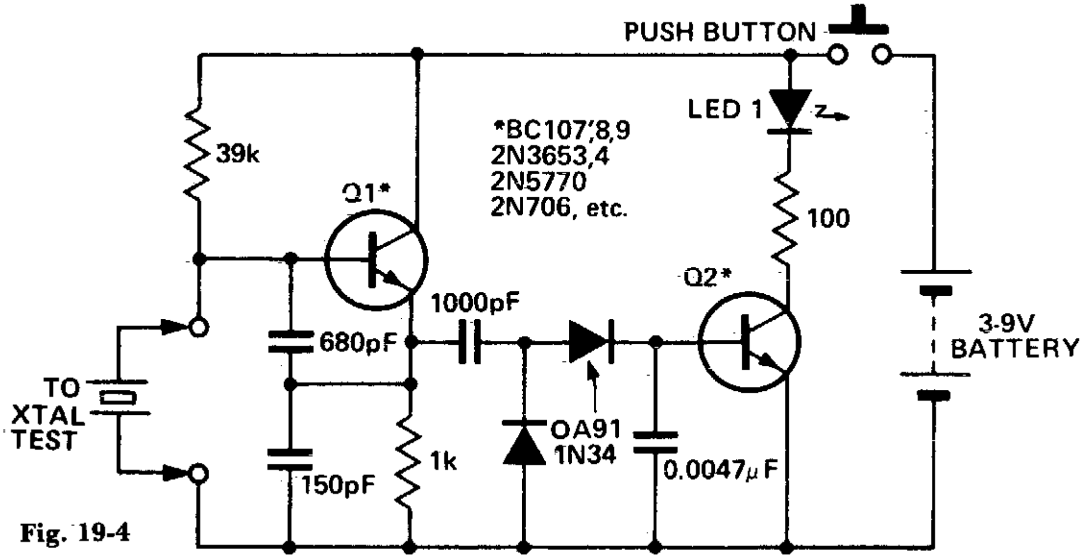
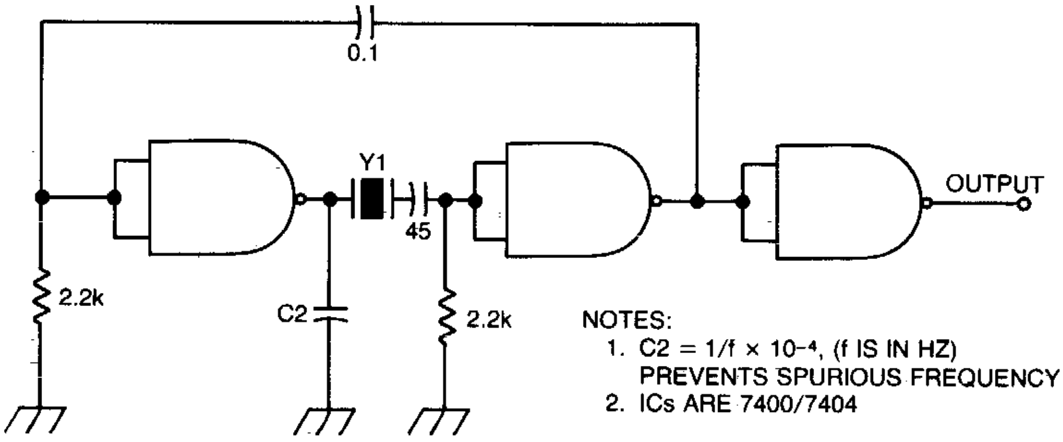

PUSH BUTTON
LED 1
*BC107,8,9
2N3653,4
2N5770
2N706, etc.
39k
Q1*
100
Q2*
1000pF
680pF
3-9V
BATTERY
TO
XTAL
TEST
OA91
1N34
1k
0.0047μF
150pF

Fig. 19-4

Circuit Notes

Use this circuit for checking fundamental HF crystals on a 'Go-No-Go' basis. An untuned Colpitts oscillator drives a voltage multiplier rectifier and a current amplifier. If the crystal oscillates, Q2 conducts and the LED lights. A3 or 6V, 40mA bulb could be substituted for the LED.

TTL OSCILLATOR FOR 1 MHz-10 MHz

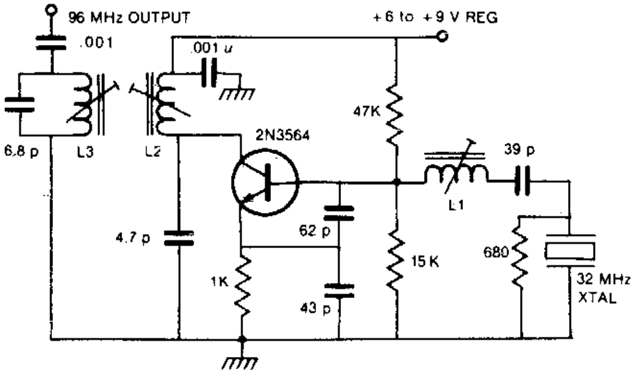

0.1
Y1
45
OUTPUT
2.2k
C2
2.2k

NOTES:

1. C2 = 1/f × 10^{-4}, (f IS IN HZ) PREVENTS SPURIOUS FREQUENCY
2. ICs ARE 7400/7404

Fig. 19-5

96 MHz CRYSTAL OSCILLATOR


96 MHz OUTPUT
.001
+6 to +9 V REG
.001 u
47K
2N3564
6.8 p
L3
L2
39 p
L1
4.7 p
62 p
15 K
680
1K
43 p
32 MHz
XTAL

L1, 4 mm former, F29 slug (Neosid AZ assembly) 30 turns .4 mm enamel wire.
L2,L3 7300 CAN TWO 722/1 FORMERS F29 SLUGS (Neosid double assembly) 12 turns .63 mm enamel wire.

Fig. 19-6

Circuit Notes

By using a crystal between 27.5 and 33 MHz, the 3rd harmonic will deliver between 82.5 and 99 MHz.

SIMPLE TTL CRYSTAL OSCILLATOR

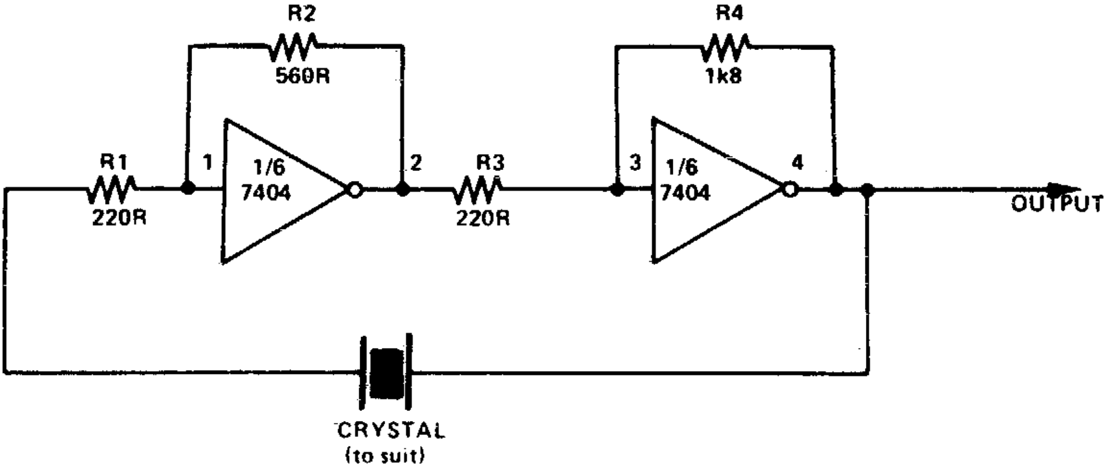

R2
560R
R4
1k8
R1
220R
1
1/6
7404
2
R3
220R
3
1/6
7404
4
OUTPUT
CRYSTAL
(to suit)

Fig. 19-7

Circuit Notes

This simple and cheap crystal oscillator comprises one third of a 7404, four resistors and a crystal. The inverters are biased into their linear regions by R1 to R4, and the crystal provides the feedback. Oscillation can only occur at the crystals fundamental frequency.

CRYSTAL OSCILLATOR

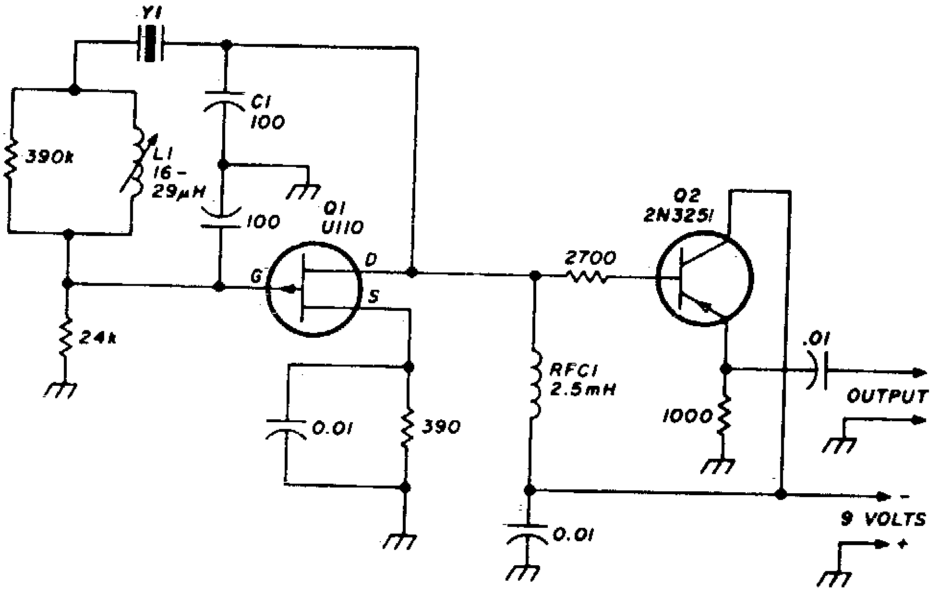

Y1
C1
100
390k
L1
16 -
29μH
100
Q1
U110
G
D
S
24k
0.01
390
Q2
2N3251
2700
RFC1
2.5mH
1000
.01
OUTPUT
0.01
-
9 VOLTS
+

Fig. 19-8

Circuit Notes

Stable VXO using 6- or 8-MHz crystals uses a capacitor and an inductor to achieve frequency pulling on either side of series resonance.

OVERTONE CRYSTAL OSCILLATOR

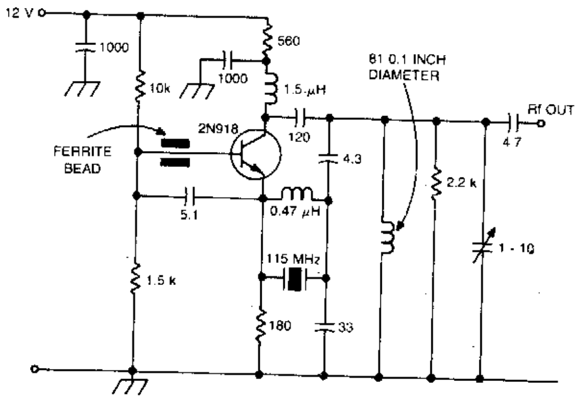

12 V
1000
10k
1000
560
1.5 μH
2N918
FERRITE
BEAD
120
4.3
81 0.1 INCH
DIAMETER
Rf OUT
4.7
2.2 k
5.1
0.47 μH
115 MHz
1.5 k
180
33
1 - 10

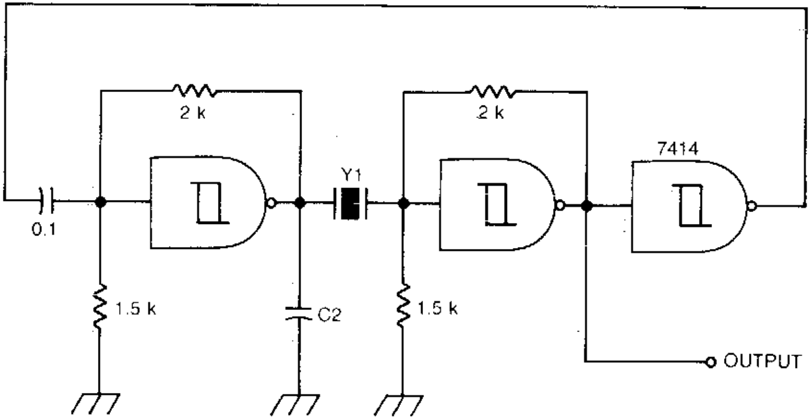
Fig. 19-9

Circuit Notes

This design is for high reliability over a wide temperature range using fifth and seventh overtone crystals. The inductor in parallel with the crystal causes antiresonance of crystal C_0 to minimize loading. This technique is commonly used with overtone crystals.

SCHMITT TRIGGER CRYSTAL OSCILLATOR

SCHMITT TRIGGER OSCILLATOR UP TO 10 MHZ


2 k
2 k
7414
Y1
0.1
1.5 k
C2
1.5 k
OUTPUT

NOTE: C2 = 1/1 × 10^{4} (f IS IN HZ) - PREVENTS SPURIOUS FREQUENCY

Fig. 19-10

Circuit Notes

A Schmitt trigger provides good squaring of the output, sometimes eliminating the need for an extra output stage.

50 MHz-150 MHz OVERTONE OSCILLATOR

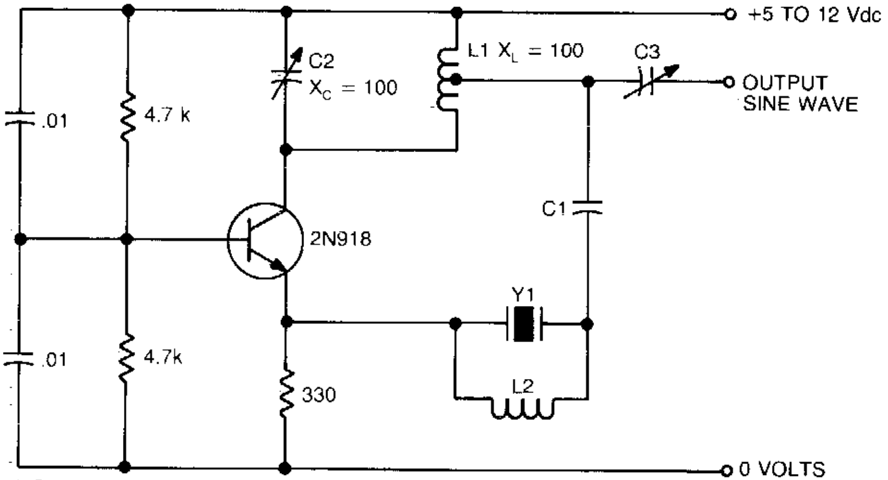

+5 TO 12 Vdc
C2
X_C = 100
L1 X_L = 100
C3
OUTPUT
SINE WAVE
.01
4.7 k
C1
2N918
Y1
.01
4.7k
330
L2
0 VOLTS

Fig. 19-11

NOTES:
1. Y1 IS AT CUT OVERTONE CRYSTAL.
2. TUNE L1 AND C2 TO OPERATING FREQUENCY.
3. L2 AND SHUNT CAPACITANCE, CO. OF CRYSTAL (APPROXIMATELY 6pF) SHOULD RESONATE TO OSCILLATOR OUTPUT FREQUENCY (L2 = .5 μH AT 90 MHZ). THIS IS NECESSARY TO TUNE OUT EFFECT OF CO.
4. C3 IS VARIED TO MATCH OUTPUT.

FIFTH-OVERTONE OSCILLATOR

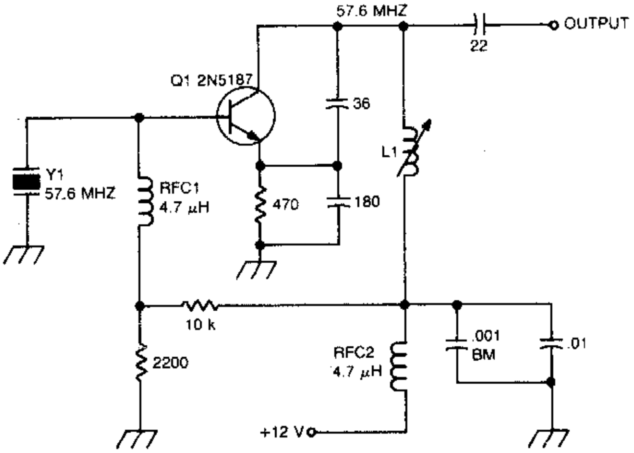

57.6 MHZ
OUTPUT
22
Q1 2N5187
36
L1
Y1
57.6 MHZ
RFC1
4.7 μH
470
180
10 k
2200
RFC2
4.7 μH
.001
BM
.01
+12 V

Fig. 19-12

Circuit Notes

This circuit isolates the crystal from the dc base supply with an rf choke for better starting characteristics.

CRYSTAL CONTROLLED BUTLER OSCILLATOR

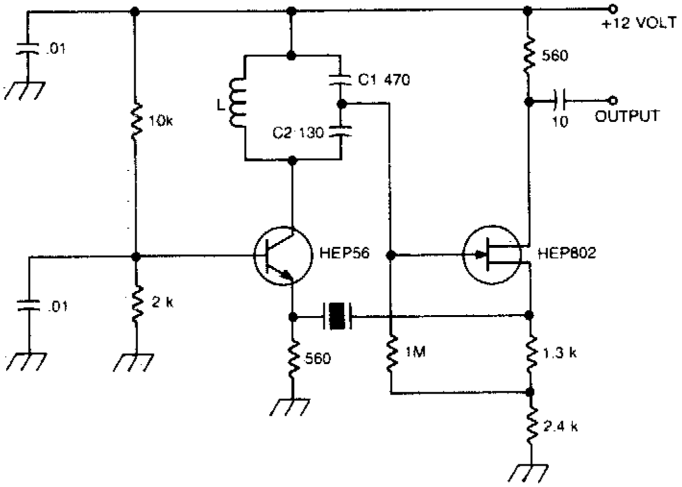

+12 VOLT
.01
560
C1 470
L
10k
C2 130
10
OUTPUT
HEP56
HEP802
.01
2 k
560
1M
1.3 k
2.4 k

Fig. 19-13

Circuit Notes

A typical Butler oscillator (20-100 MHz) uses an FET in the second stage; the circuit is not reliable with two bipolars. Sometimes two FETs are used. Frequency is determined by LC values.

OVERTONE OSCILLATOR WITH CRYSTAL SWITCHING

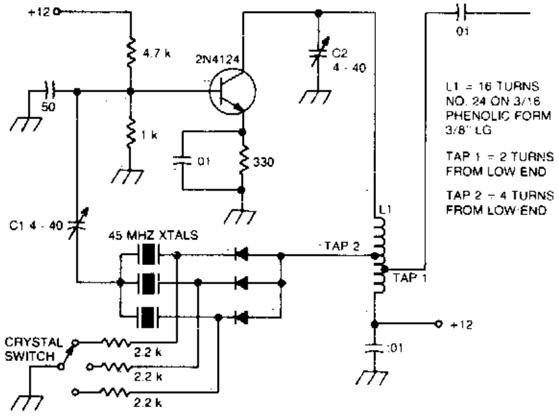

+12
4.7 k
2N4124
C2
4 - 40
50
1 k
.01
330
C1 4 - 40
45 MHZ XTALS
TAP 2
L1
TAP 1
+12
.01
CRYSTAL SWITCH
2.2 k
2.2 k
2.2 k
.01
L1 = 16 TURNS NO. 24 ON 3/16 PHENOLIC FORM 3/8" LG
TAP 1 = 2 TURNS FROM LOW END
TAP 2 = 4 TURNS FROM LOW END

Fig. 19-14

Circuit Notes

The large inductive phase shift of L1 is compensated for by C1. Overtone crystals have very narrow bandwidth; therefore, the trimmer has a smaller effect than for fundamental-mode operation.

CRYSTAL OSCILLATOR

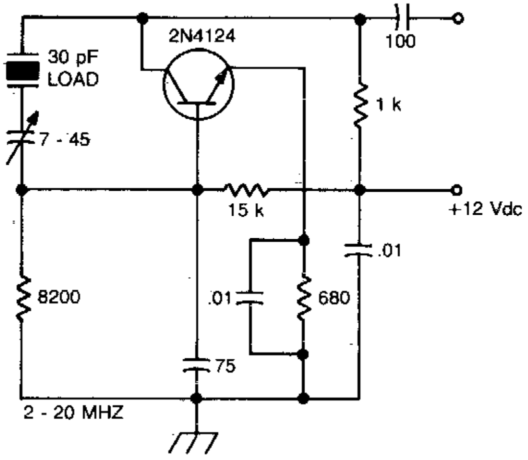

2N4124
30 pF
LOAD
7 - 45
100
1 k
15 k
+12 Vdc
.01
8200
.01
680
75
2 - 20 MHZ

Fig. 19-15

Circuit Notes

The crystal is in a feedback circuit from collector to base. A trimmer capacitor in series shifts the point on the reactance curve where the crystal operates, thus providing a frequency trim. The capacitor has a negative reactance so the crystal is shifted to operate in the positive reactance region.

CRYSTAL OSCILLATOR/DOUBLER

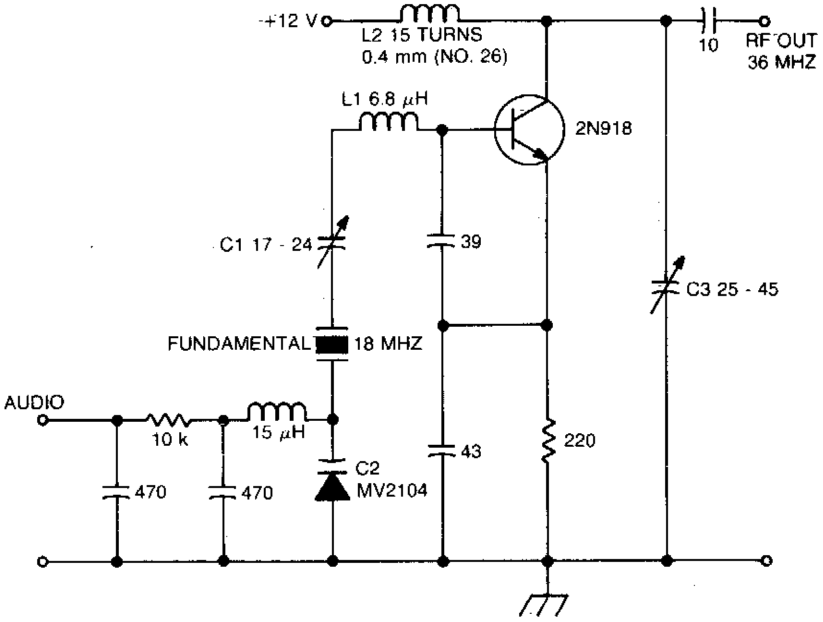

+12 V
L2 15 TURNS
0.4 mm (NO. 26)
10
RF OUT
36 MHZ
L1 6.8 μH
2N918
C1 17 - 24
39
C3 25 - 45
FUNDAMENTAL
18 MHZ
AUDIO
10 k
15 μH
220
43
C2
MV2104
470
470

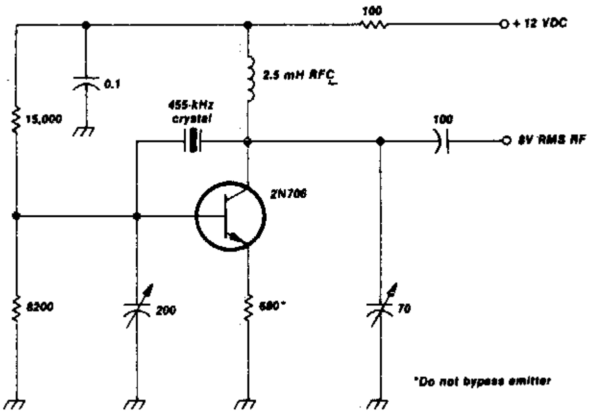
Fig. 19-16

Circuit Notes

The crystal operates into a complex load at series resonance. L1, C1, and C2 balance the crystal at zero reactance. Capacitor C1 fine-tunes the center frequency. Tank circuit L2, C3 doubles the output frequency the circuit operates as an FM oscillator-doubler.

LOW-FREQUENCY CRYSTAL OSCILLATOR

*Except as indicated, decimal values of capacitance are in microfarads (μF); others are in picofarads (pF); resistances are in ohms.
k = 1,000 M = 1,000,000*


100
+ 12 VDC
0.1
2.5 mH RFC
15,000
455-kHz crystal
100
8V RMS RF
2N706
8200
200
680*
70

*Do not bypass emitter

Fig. 19-17

Circuit Notes

This crystal-oscillator circuit uses a 455-kHz crystal.

CRYSTAL OSCILLATOR

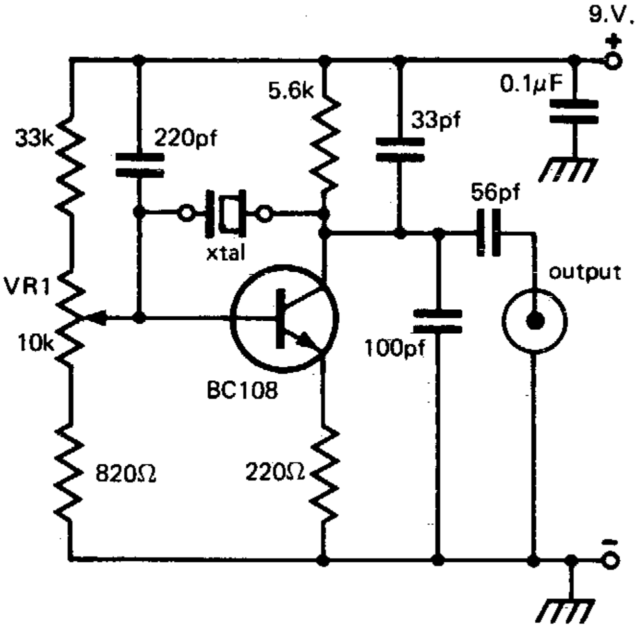

9.V.
+
0.1µF
5.6k
33pf
33k
220pf
56pf
xtal
output
VR1
10k
BC108
100pf
820Ω
220Ω
-

Fig. 19-18

Circuit Notes

This circuit provides reliable oscillation and an output close to one volt peak-to-peak. Power consumption is around 1 mA from a nine volt supply.

100 kHz CRYSTAL CALIBRATOR

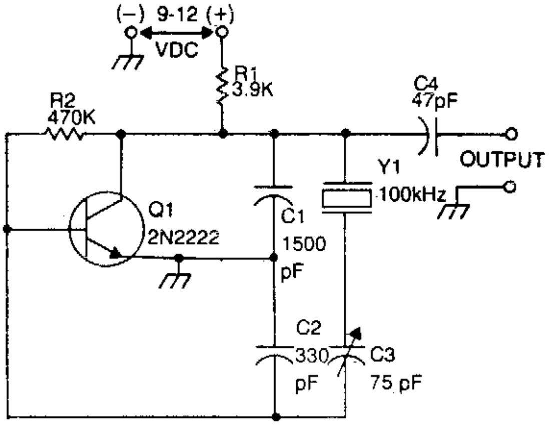
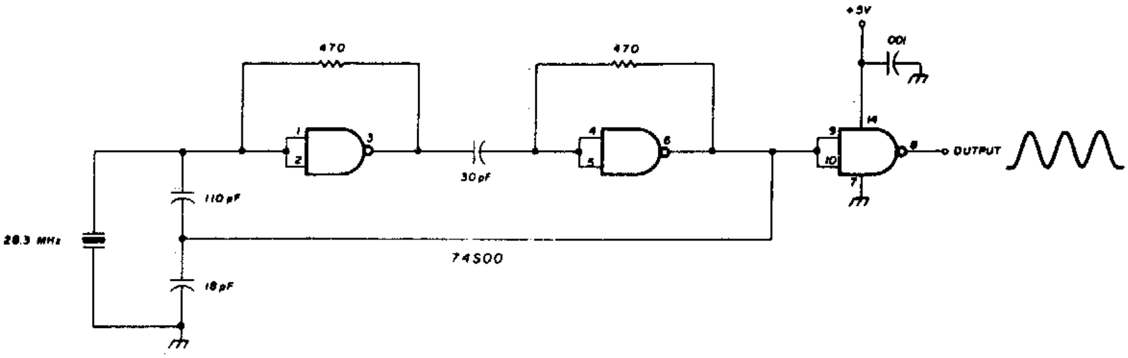

(−) 9-12 (+)
VDC
R1
3.9K
R2
470K
C4
47pF
OUTPUT
Y1
100kHz
Q1
2N2222
C1
1500
pF
C2
330
pF
C3
75 pF

Fig. 19-19

Circuit Notes

This circuit is often used by amateur radio operations, shortwave listeners, and other operators of shortwave receivers to calibrate the dial pointer. The oscillator operates at a fundamental frequency of 100 kHz, and the harmonics are used to locate points on the shortwave dial, provided that the output of the calibrator is coupled to the antenna circuit of the receiver. The crystal shunts the feedback voltage divider, and is in series with a variable capacitor (C3) that is used to set the actual operating frequency of the calibrator.

THIRD-OVERTONE CRYSTAL OSCILLATOR



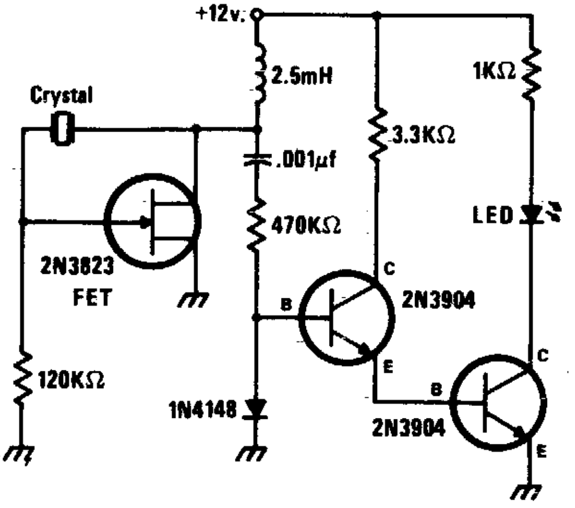
470
470
+5V
.001
1
2
3
4
5
6
9
10
14
8
7
OUTPUT
30pF
110pF
28.3 MHz
74S00
18pF

Fig. 19-20

Circuit Notes

This circuit uses a 74S00 Schottky TTL gate; no inductors are required.

CRYSTAL CHECKER



+12v.
2.5mH
1KΩ
Crystal
3.3KΩ
.001μf
LED
470KΩ
2N3823
FET
C
B
2N3904
E
C
B
120KΩ
1N4148
2N3904
E

Circuit Notes

This circuit is a simple Pierce oscillator with an LED go/no go display. Checker works best with crystals having fundamental frequencies in the seven to eight megahertz range.

Fig. 19-21

CMOS CRYSTAL OSCILLATOR

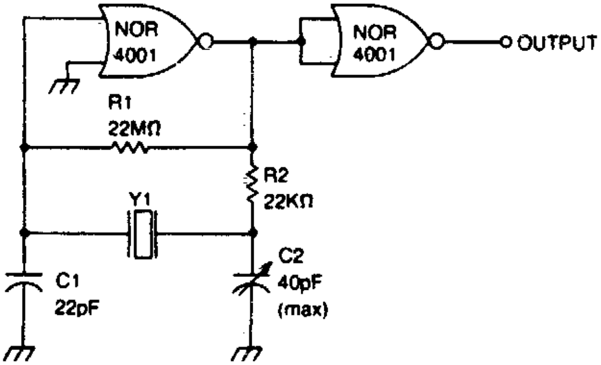
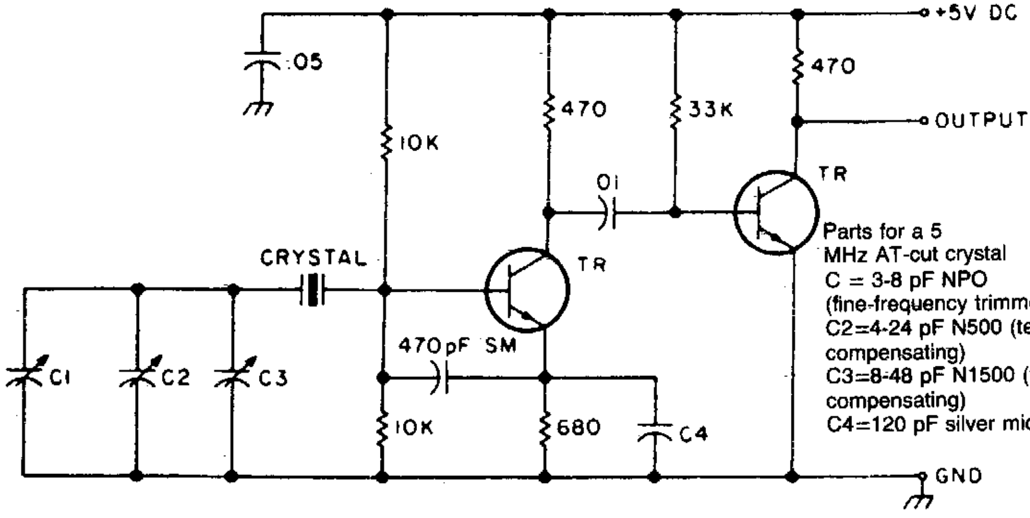

NOR
4001
NOR
4001
OUTPUT
R1
22MΩ
R2
22KΩ
Y1
C1
22pF
C2
40pF
(max)

Fig. 19-22

Circuit Notes

This circuit has a frequency range of 0.5 MHz to 2.0 MHz. Frequency can be adjusted to a precise value with trimmer capacitor C2. The second NOR gate serves as an output buffer.

TEMPERATURE-COMPENSATED CRYSTAL OSCILLATOR


+5V DC
.05
10K
470
33K
470
OUTPUT
.01
TR
CRYSTAL
TR
470pF SM
C1
C2
C3
10K
680
C4
GND

Parts for a 5 MHz AT-cut crystal
C = 3-8 pF NPO (fine-frequency trimmer)
C2=4-24 pF N500 (temperature compensating)
C3=8-48 pF N1500 (temperature compensating)
C4=120 pF silver mica

Fig. 19-23

Circuit Notes

Two different negative-coefficient capacitors are blended to produce the desired change in capacitance to counteract or compensate for the decrease in frequency of the "normal" AT-cut characteristics.

CRYSTAL-CONTROLLED, TRANSISTOR OSCILLATOR

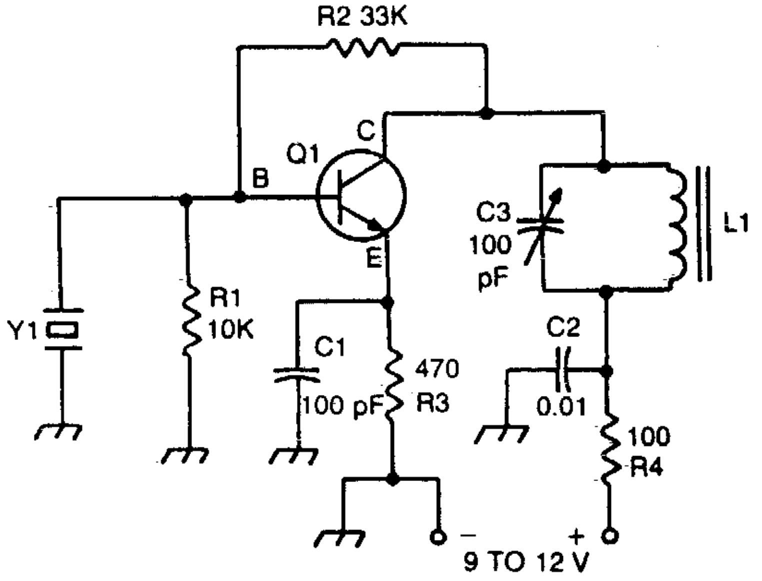

R2 33K
C
Q1
B
E
C3
100
pF
L1
Y1
R1
10K
C1
100 pF
470
R3
C2
0.01
100
R4
–
+
9 TO 12 V

Fig. 19-24

PIERCE HARMONIC OSCILLATOR (20 MHz)

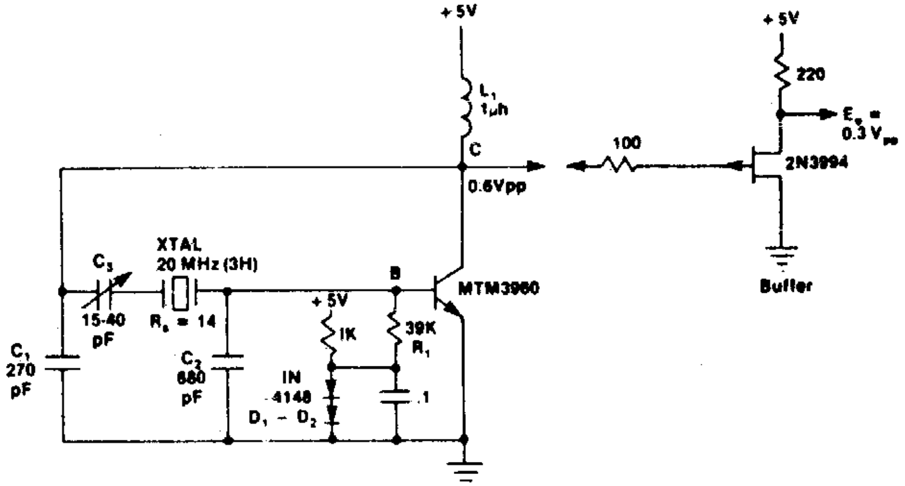

+5V
L1
1μh
C
0.6Vpp
100
+5V
220
E_o =
0.3 V_{pp}
2N3994
Buffer
XTAL
20 MHz (3H)
C_3
15-40
pF
R_s = 14
B
MTM3960
+5V
1K
39K
R_1
C_1
270
pF
C_2
680
pF
IN
4148
D_1 – D_2
.1

Fig. 19-25

Circuit Notes

This circuit has excellent short term frequency stability because the external load tied across the crystal is mostly capacitive rather than resistive, giving the crystal a high in-circuit Q.

COLPITTS HARMONIC OSCILLATOR (100 MHz)

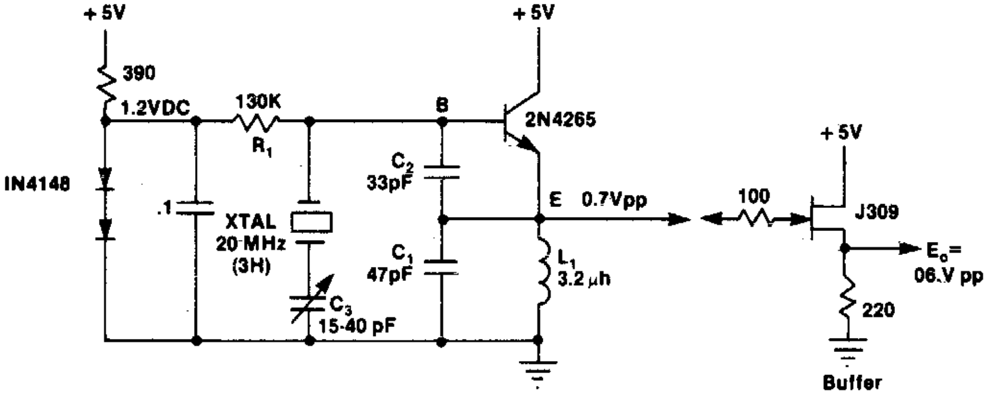
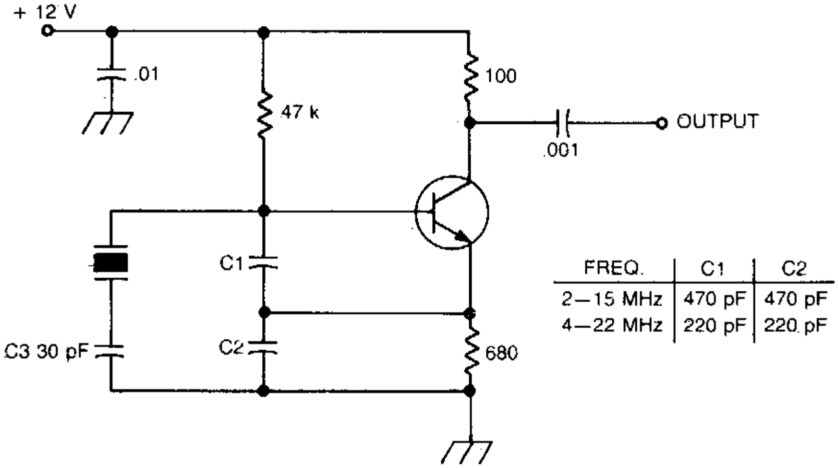

+5V
390
1.2VDC
130K
R1
B
+5V
2N4265
IN4148
.1
XTAL
20 MHz
(3H)
C2
33pF
C1
47pF
C3
15-40 pF
E 0.7Vpp
L1
3.2 μh
100
+5V
J309
Eo=
06.V pp
220
Buffer

Fig. 19-26

Circuit Notes

L1C1 are selected to be resonant at a frequency below the desired crystal harmonic but above the crystal's next lower odd harmonic. C2 should have a value of 30-70 pF, independent of the oscillation frequency. There is no requirement for any specific ratio of C1/C2, but practical harmonic circuits seem to work best when C1 is approximately 1-3 times the value of C2. Diodes D1-D3 provide a simple regulated bias supply. The resistance of R1 should be as high as possible, as it affects the crystal's in-circuit Q.

INTERNATIONAL CRYSTAL OF-1 LO OSCILLATOR


+ 12 V
.01
47 k
100
OUTPUT
.001
C1
C2
C3 30 pF
680

FREQ.	C1	C2
2—15 MHz	470 pF	470 pF
4—22 MHz	220 pF	220 pF

Fig. 19-27

Circuit Notes

International Crystal OF-1 LO oscillator circuit for fundamental-mode crystals.

BUTLER EMITTER FOLLOWER OSCILLATOR (100 MHz)

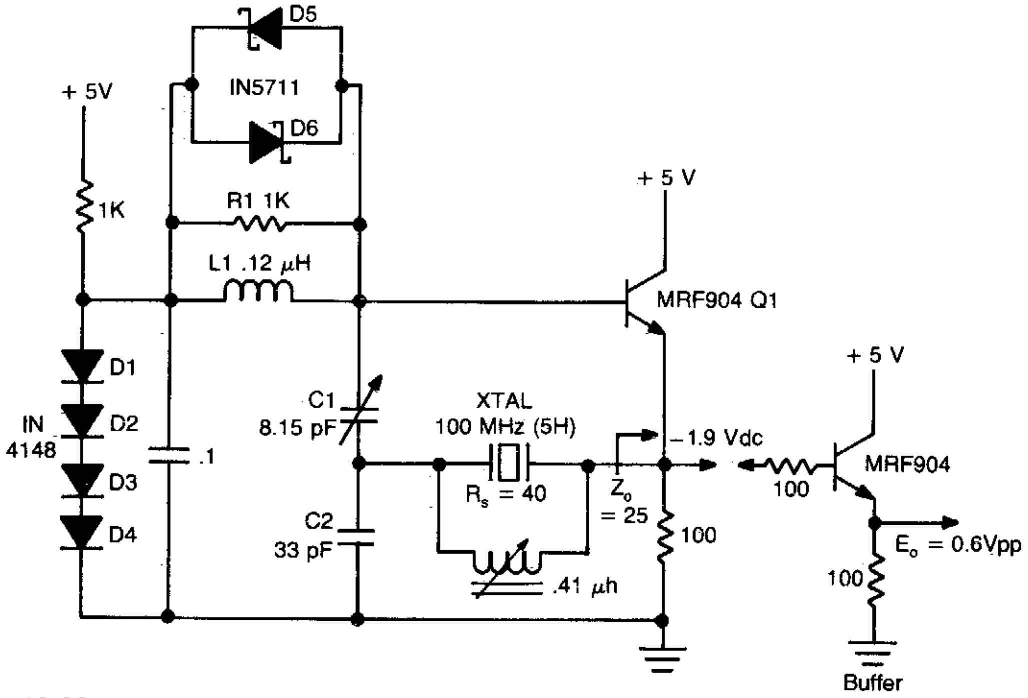
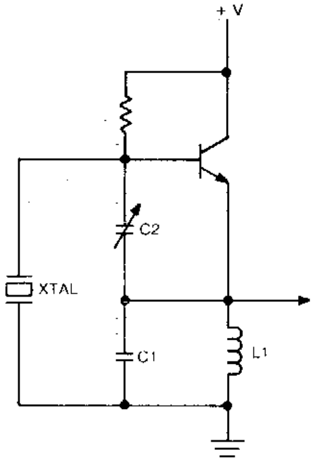

D5
IN5711
D6
+ 5V
1K
R1 1K
L1 .12 µH
D1
IN
4148
D2
D3
D4
.1
C1
8.15 pF
C2
33 pF
XTAL
100 MHz (5H)
R_s = 40
Z_o
= 25
.41 µh
+ 5 V
MRF904 Q1
−1.9 Vdc
100
+ 5 V
100
MRF904
E_o = 0.6Vpp
100
Buffer

Fig. 19-28

Circuit Notes

This circuit has good performance without any parasitics because emitter follower amplifier has a gain of only one with built-in negative feedback to stabilize its gain.

COLPITTS HARMONIC OSCILLATOR (BASIC CIRCUIT)

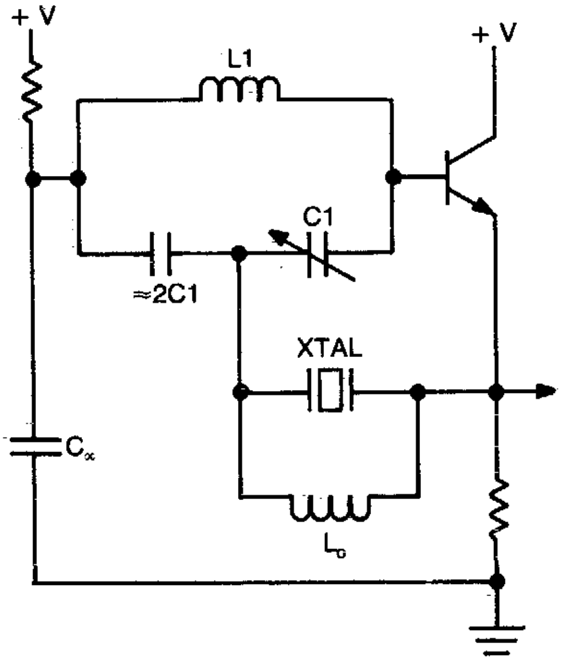

+ V
C2
XTAL
C1
L1

Circuit Notes

This circuit operates 30-200 ppm above series resonance. Physically simple, but analytically complex. It is inexpensive with fair frequency stability.

Fig. 19-29

BUTLER EMITTER FOLLOWER OSCILLATOR (BASIC CIRCUIT)

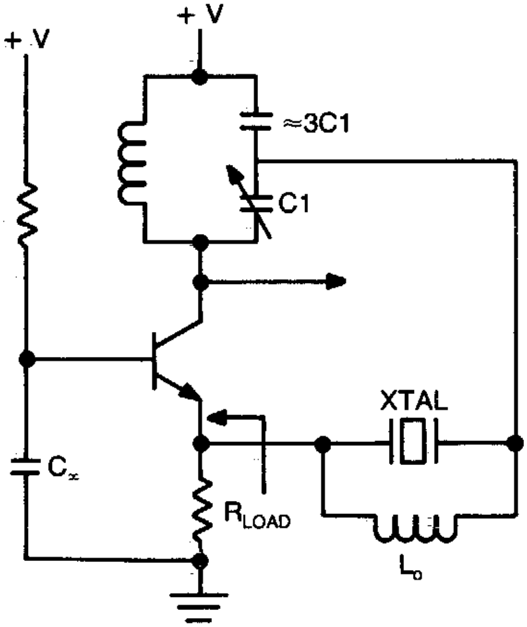

+V
L1
+V
C1
≈2C1
XTAL
C_∞
L_o

Circuit Notes

This circuit operates at or near series resonance. It is a good circuit design with no parasitics. It is easy to tune with good frequency stability.

Fig. 19-30

BUTLER COMMON BASE OSCILLATOR (BASIC CIRCUIT)

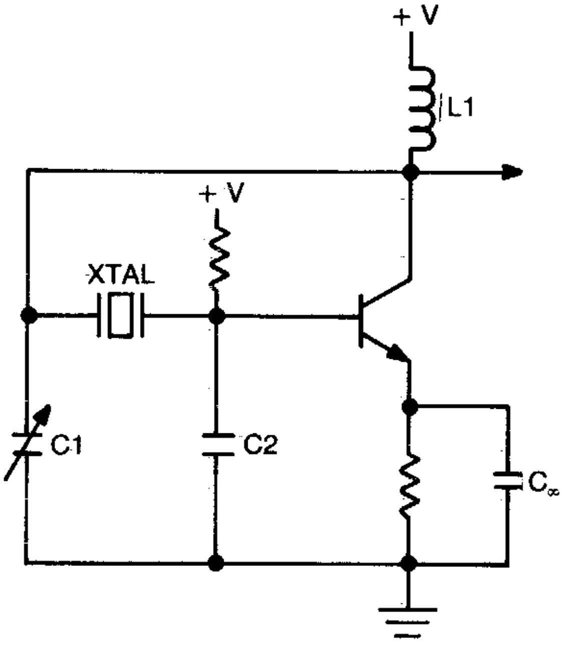

+V
+V
≈3C1
C1
XTAL
C_∞
R_{LOAD}
L_o

Circuit Notes

This circuit operates at or near series resonance. It has fair to poor circuit design with parasitics, touch to tune, and fair frequency stability.

Fig. 19-31

PIERCE HARMONIC OSCILLATOR (BASIC CIRCUIT)



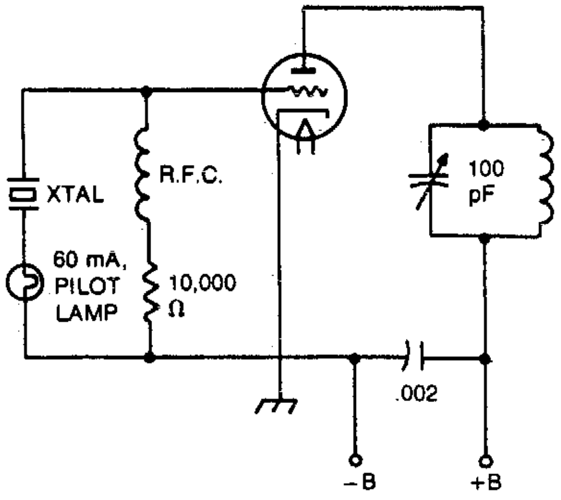
+ V
L1
+ V
XTAL
C1
C2
C_∞

Circuit Notes

This circuit operates 10-40 ppm above series resonance. It is a good circuit design with good to very good frequency stability.

Fig. 19-32

TUBE-TYPE CRYSTAL OSCILLATOR



XTAL
R.F.C.
60 mA, PILOT LAMP
10,000 Ω
100 pF
.002
−B
+B

Circuit Notes

The pilot lamp limits current to prevent damage to the crystal.

Fig. 19-33

PRECISION CLOCK GENERATOR

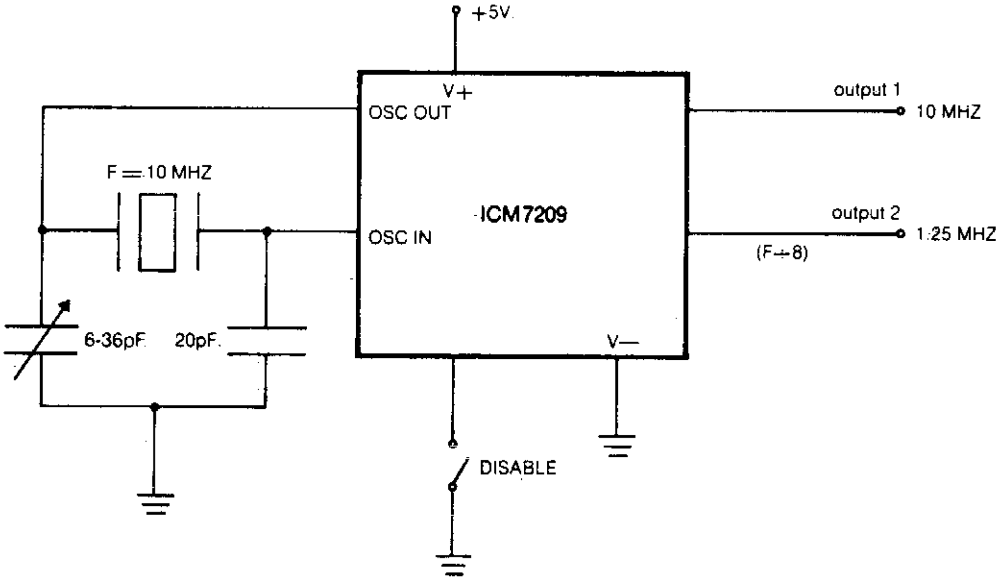
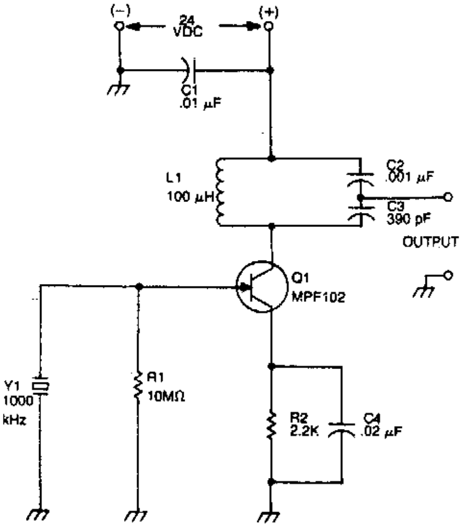

+5V
V+
OSC OUT
output 1
10 MHZ
F = 10 MHZ
ICM7209
OSC IN
output 2
1.25 MHZ
(F÷8)
6-36pF
20pF
V−
DISABLE

Fig. 19-34

Circuit Notes

The CMOS IC directly drives 5 TTL loads from either of 2 buffered outputs. The device operates to 10 MHz and is bipolar, MOS, and CMOS compatible.

MILLER OSCILLATOR (CRYSTAL CONTROLLED)

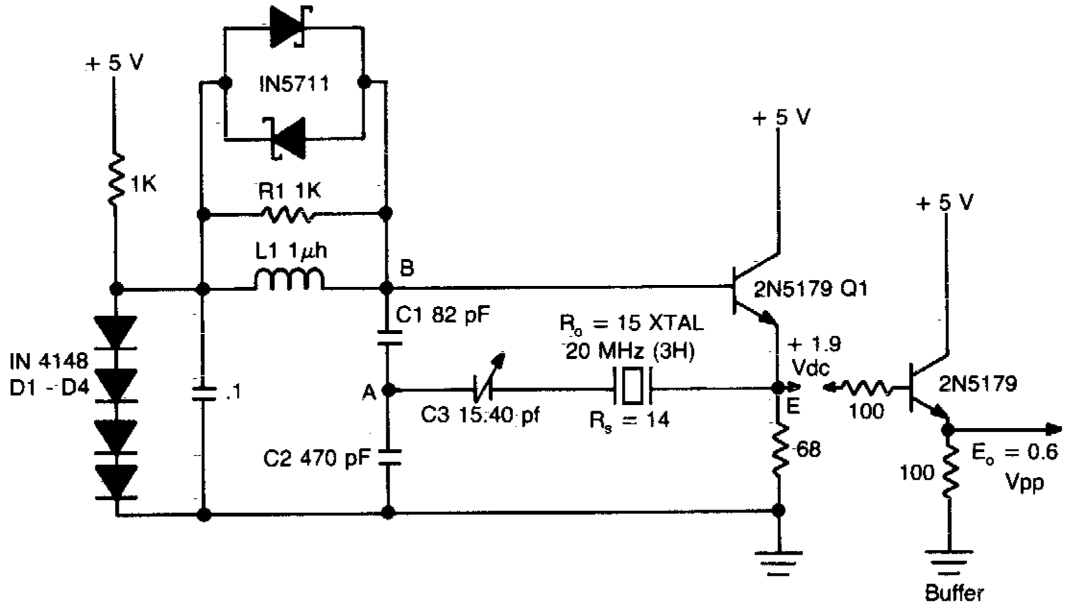

(−)
(+)
24 VDC
C1
.01 µF
L1
100 µH
C2
.001 µF
C3
390 pF
OUTPUT
Q1
MPF102
Y1
1000 kHz
R1
10MΩ
R2
2.2K
C4
.02 µF

Circuit Notes

The drain of the JFET Miller oscillator is tuned to the resonant frequency of the crystal by an LC tank circuit.

Fig. 19-35

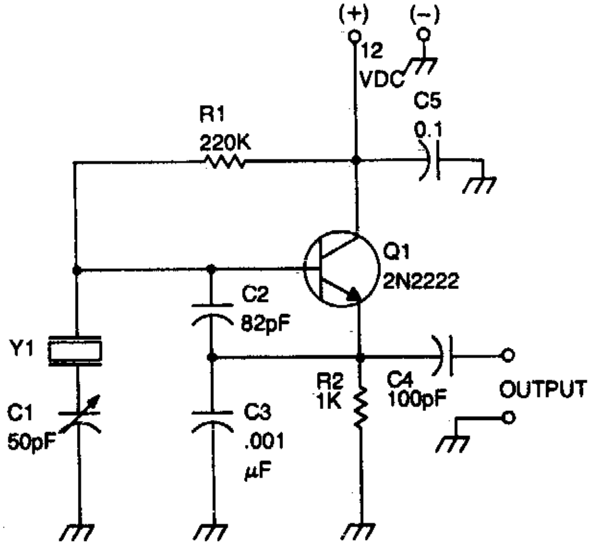
BUTLER EMITTER FOLLOWER OSCILLATOR (20 MHz)



+ 5 V
1K
IN5711
R1 1K
L1 1μh
B
IN 4148
D1 - D4
.1
C1 82 pF
A
C2 470 pF
C3 15-40 pf
R_o = 15 XTAL
20 MHz (3H)
R_s = 14
+ 5 V
2N5179 Q1
+ 1.9
Vdc
E
68
100
+ 5 V
2N5179
100
E_o = 0.6
Vpp
Buffer

Fig. 19-36

COLPITTS OSCILLATOR



(+)
(−)
12
VDC
R1
220K
C5
0.1
Q1
2N2222
C2
82pF
Y1
C1
50pF
C3
.001
μF
R2
1K
C4
100pF
OUTPUT

Circuit Notes

This circuit will operate with fundamental-mode crystals in the range of 1 MHz to 20 MHz. Feedback is controlled by capacitor voltage divider C2/C3. The rf voltage across the emitter resistor provides the basic feedback signal.

Fig. 19-37

CRYSTAL-CONTROLLED OSCILLATOR

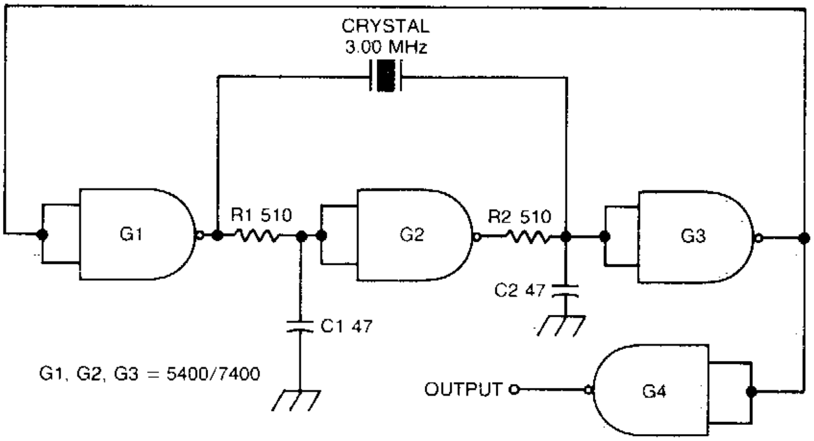

CRYSTAL
3.00 MHz
G1
R1 510
G2
R2 510
G3
C1 47
C2 47
G1, G2, G3 = 5400/7400
OUTPUT
G4

Fig. 19-38

Circuit Notes

This circuit oscillates without the crystal. With the crystal in the circuit, the frequency will be that of the crystal. The circuit has good starting characteristics even with the poorest crystals.

PIERCE OSCILLATOR

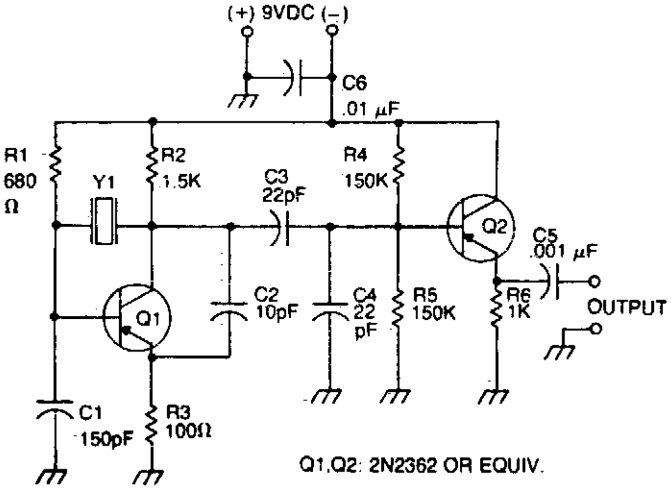
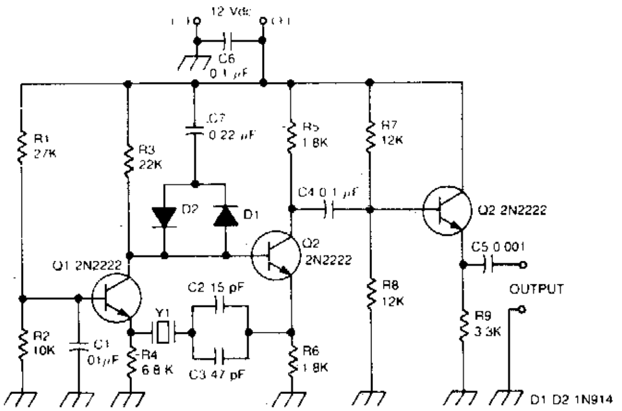

(+) 9VDC (–)
C6
.01 μF
R1
680
Ω
Y1
R2
1.5K
C3
22pF
R4
150K
Q2
C5
.001 μF
Q1
C2
10pF
C4
22
pF
R5
150K
R6
1K
OUTPUT
C1
150pF
R3
100Ω
Q1,Q2: 2N2362 OR EQUIV.

Fig. 19-39

Circuit Notes

The oscillator transistor is Q1, and the crystal is placed between the collector and base. Feedback is improved by the use of the collector-emitter capacitor C2. Transistor Q2 is used as an output buffer.

BUTLER APERIODIC OSCILLATOR



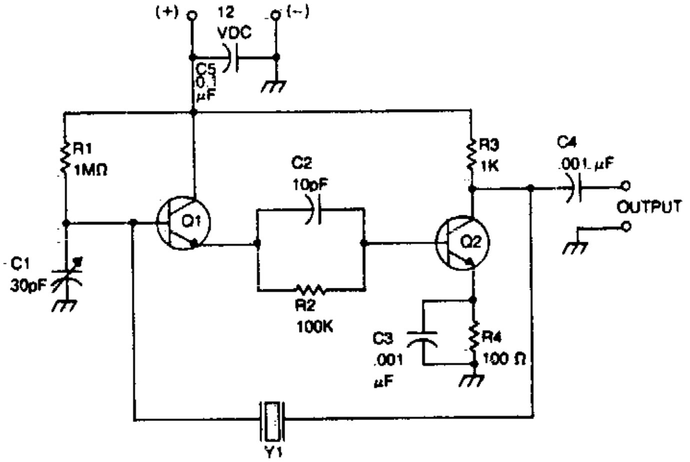
12 Vdc
(−)
(+)
C6
0.1 μF
C7
0.22 μF
R1
27K
R3
22K
R5
1.8K
R7
12K
C4 0.1 μF
Q2 2N2222
D2
D1
Q2
2N2222
C5 0.001
OUTPUT
Q1 2N2222
C2 15 pF
R8
12K
Y1
R2
10K
C1
.01 μF
R4
6.8 K
C3 47 pF
R6
1.8K
R9
3.3K
D1 D2 1N914

Fig. 19-40

Circuit Notes

This circuit works well in the range of 50 kHz to 500 kHz. Slight component modifications are needed for higher frequency operation. For operation over 3000 kHz, select a transistor that provides moderate gain (in the 60 to 150 range) at the frequency of operation and a gain-bandwidth product of at least 100 MHz.

PARALLEL-MODE APERIODIC CRYSTAL OSCILLATOR



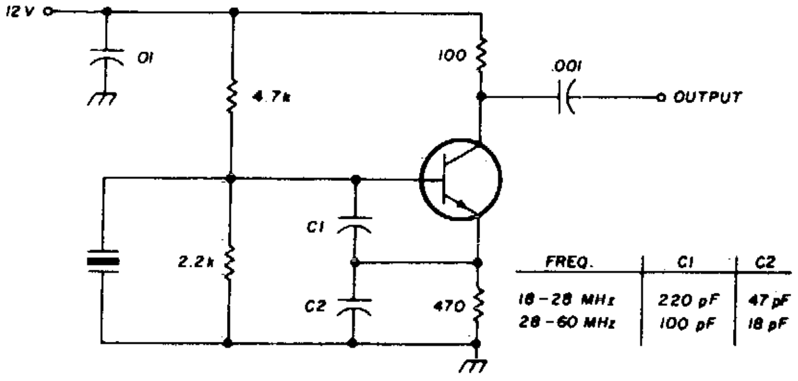
(+)
12
VDC
(−)
C5
0.1
μF
R1
1MΩ
C2
10pF
R3
1K
C4
.001 μF
OUTPUT
Q1
Q2
C1
30pF
R2
100K
C3
.001
μF
R4
100 Ω
Y1

Fig. 19-41

Circuit Notes

The crystal is placed between the collector of the output stage and the base of the input stage. The frequency of oscillation can be set to a precise value with trimmer capacitor C1. The range of operation for this circuit is 500 kHz to 10 MHz. Extend the range downward (100 kHz) by increasing the value of C1 to 75 pF and increasing the value of C2 to 22pF.

INTERNATIONAL CRYSTAL OF-1 HI OSCILLATOR


12V
.01
4.7k
100
.001
OUTPUT
C1
C2
2.2k
470

FREQ.	C1	C2
18-28 MHz	220 pF	47 pF
28-60 MHz	100 pF	18 pF

Fig. 19-42

Circuit Notes

International Crystal OF-1 HI oscillator circuit for third-overtone crystals. The circuit does not require inductors.

STANDARD CRYSTAL OSCILLATOR FOR 1 MHz

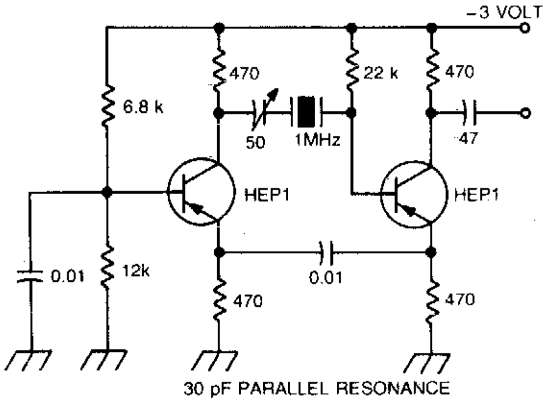

-3 VOLT
6.8 k
470
22 k
470
50
1MHz
47
HEP1
HEP1
0.01
12k
0.01
470
470
30 pF PARALLEL RESONANCE

Fig. 19-43

TTL-COMPATIBLE CRYSTAL OSCILLATOR

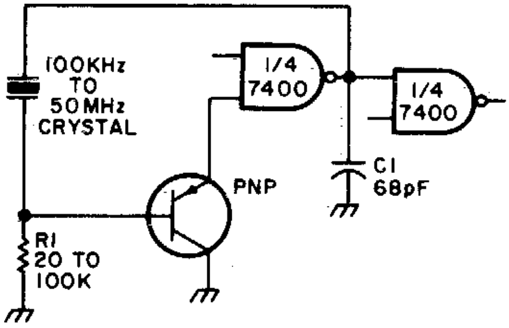
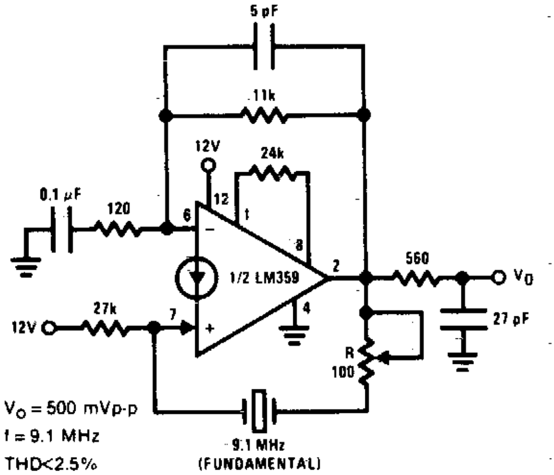

100KHz TO 50MHz CRYSTAL
1/4 7400
1/4 7400
C1 68pF
PNP
R1 20 TO 100K

Fig. 19-44

Circuit Notes

Adjust R1 for about 2 volts at the output of the first gate. Adjust C1 for best output.

CRYSTAL CONTROLLED SINE WAVE OSCILLATOR


5 pF
11k
12V
24k
0.1 μF
120
6
12
1
8
1/2 LM359
2
560
V₀
4
27 pF
27k
12V
7
+
−
R
100
9.1 MHz
(FUNDAMENTAL)

V_O = 500 mVp-p
f = 9.1 MHz
THD<2.5%

Fig. 19-45

STABLE LOW FREQUENCY CRYSTAL OSCILLATOR

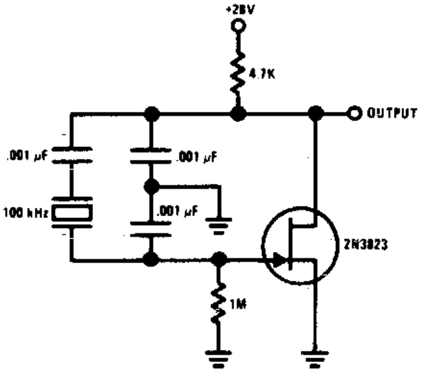

+28V
4.7K
OUTPUT
.001 μF
.001 μF
100 kHz
.001 μF
2N3823
1M

Fig. 19-47

Circuit Notes

This Colpitts-crystal oscillator is ideal for low frequency crystal oscillator circuits. Excellent stability is assured because the 2N3823 JFET circuit loading does not vary with temperature.

CRYSTAL OSCILLATOR

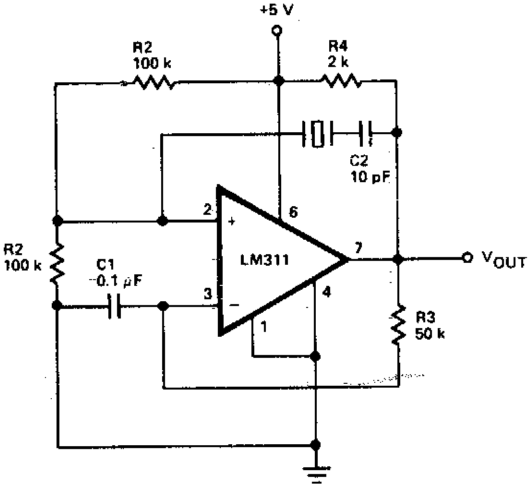

+5 V
R2
100 k
R4
2 k
C2
10 pF
2
+
6
LM311
7
V_{OUT}
R2
100 k
C1
0.1 μF
3
−
4
1
R3
50 k

Fig. 19-46

JFET PIERCE CRYSTAL OSCILLATOR

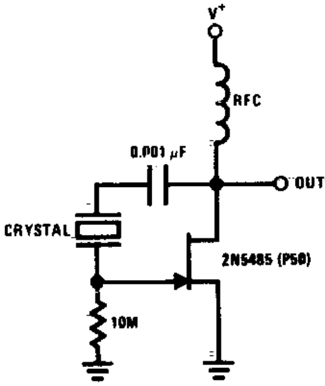
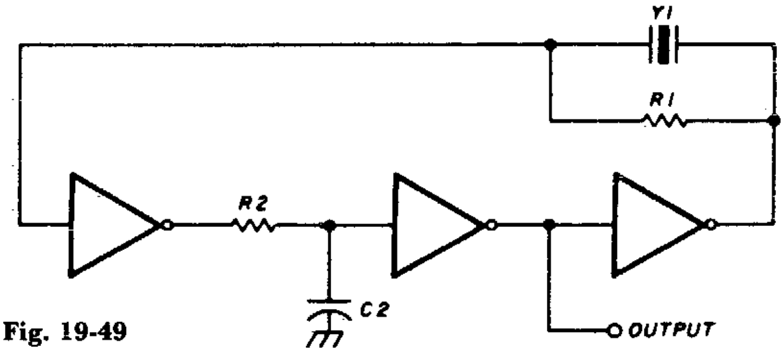

V⁺
RFC
0.001 μF
OUTPUT
CRYSTAL
2N5485 (P50)
10M

Fig. 19-48

Circuit Notes

The JFET Pierce crystal oscillator allows a wide frequency range of crystals to be used without circuit modification. Since the JFET gate does not load the crystal, good Q is maintained, thus insuring good frequency stability.

CMOS OSCILLATOR-1 MHz-4 MHz


Y1
R1
R2
C2
OUTPUT

NOTES:
1. 1M < R1 < 5M
2. SELECT R2 AND C2 TO PREVENT SPURIOUS FREQUENCIES
3. ICs ARE 74C04 OR EQUIVALENT

Fig. 19-49

PIERCE HARMONIC OSCILLATOR (100 MHz)

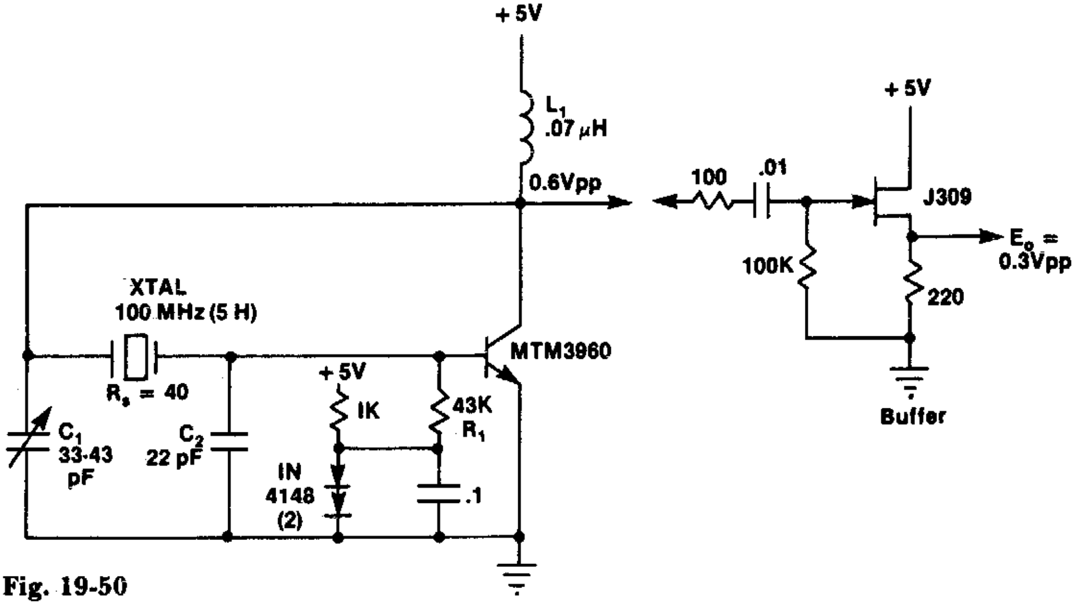

+5V
L1
.07 μH
0.6Vpp
100
.01
+5V
J309
100K
220
E0 = 0.3Vpp
Buffer
XTAL
100 MHz (5 H)
Rs = 40
MTM3960
+5V
IK
43K
R1
C1
33-43 pF
C2
22 pF
IN 4148 (2)
.1

Fig. 19-50

Circuit Notes

The output resistance of the transistor's collector, together with the effective value of C1, provides an RC phase lag of 30-50°. The crystal normally oscillates slightly above series resonance, where it is both resistive and inductive. Above series resonance, the crystal's internal impedance (resistive and inductive) together with C2 provides an RLC phase lag of 130-150°. The transistor inverts the signal, providing a total of 360° of phase shift around the loop. Inductor L1 is selected to resonate with C1 at a frequency between the crystal's desired harmonic and its next lower odd harmonic. Inductor L1 offsets part of the negative reactance of C1 at the oscillation frequency.

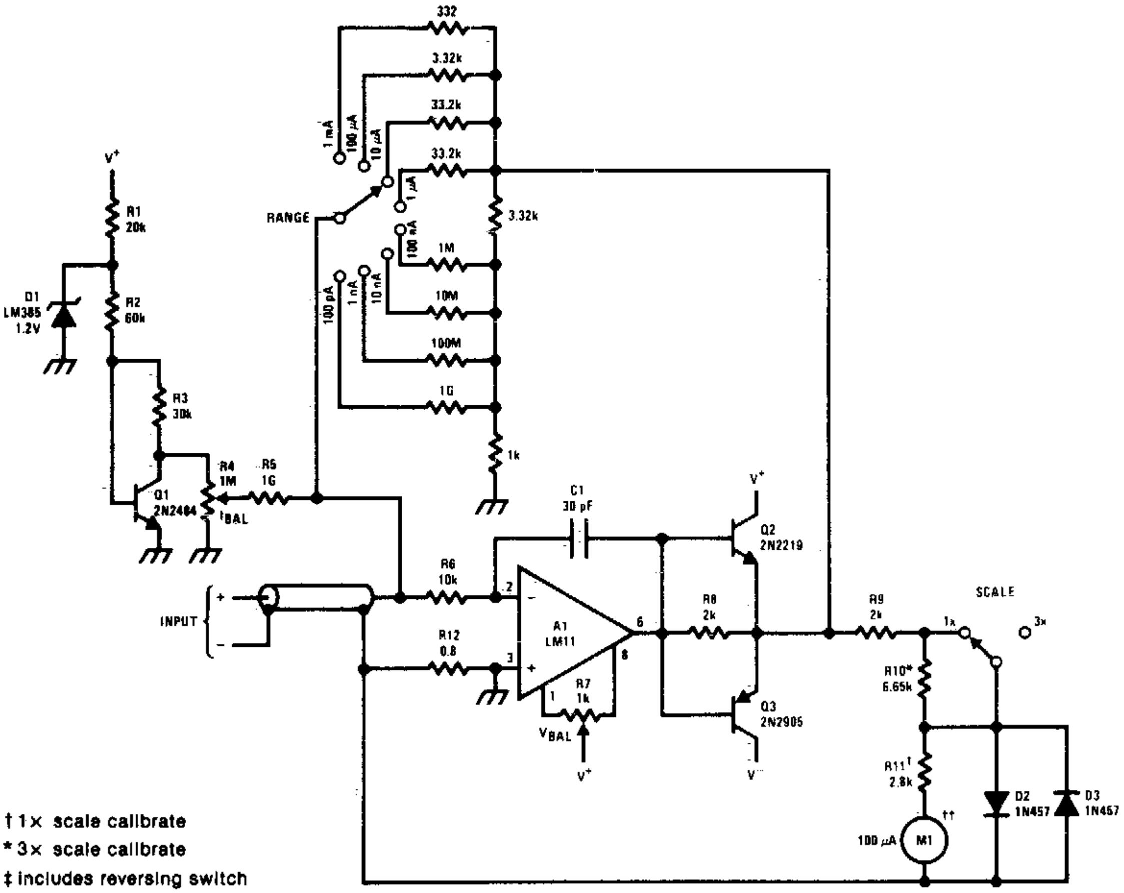
20

Current Measuring Circuits

The sources of the following circuits are contained in the Sources section beginning on page 730. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

Ammeter
Pico Ammeter
Nano Ammeter
Nanoampere Sensing Circuit with 100 Megohm Input Impedance
Current Monitor

AMMETER


332
3.32k
33.2k
33.2k
3.32k
1M
10M
100M
1G
1k
1 mA
100 μA
10 μA
1 μA
100 nA
10 nA
1 nA
100 pA
RANGE
V+
R1
20k
D1
LM385
1.2V
R2
60k
R3
30k
Q1
2N2484
R4
1M
R5
1G
IBAL
C1
30 pF
V+
Q2
2N2219
R6
10k
INPUT
+
−
R12
0.8
A1
LM11
2
3
6
8
1
R7
1k
VBAL
V+
R8
2k
Q3
2N2905
V−
R9
2k
SCALE
1x
3x
R10*
6.65k
R11†
2.8k
100 μA
M1
††
D2
1N457
D3
1N457

† 1x scale calibrate
* 3x scale calibrate
‡ includes reversing switch

Fig. 20-1

Circuit Notes

Current meter ranges from 100 pA to 3 mA full scale. Voltage across input is 100 μV at lower ranges rising to 3 mV at 3 mA. The buffers on the op amp are to remove ambiguity with high-current overload. The output can also drive a DVM or a DPM.

PICO AMMETER

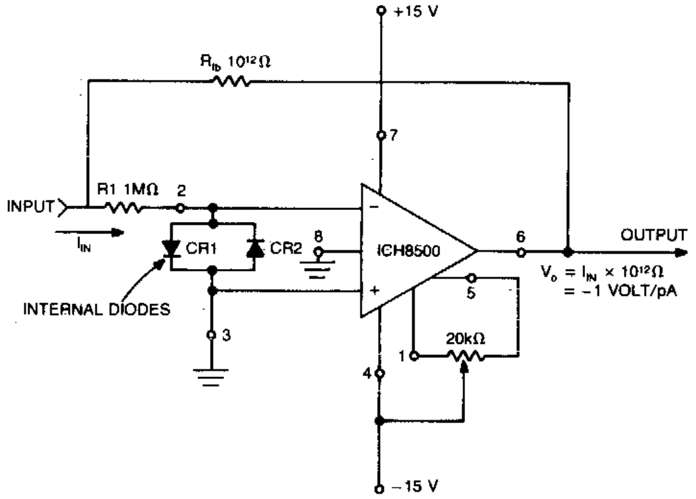
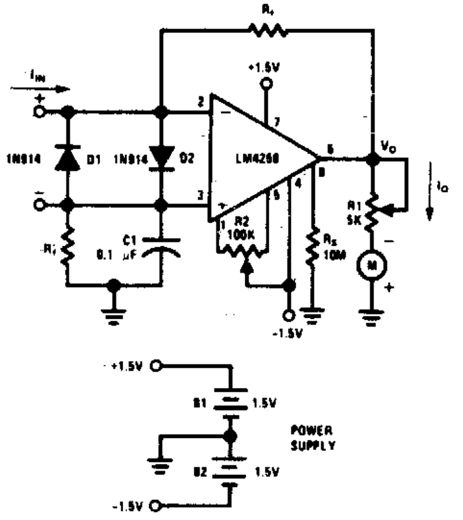

+15 V
R_{fb} $10^{12}\Omega$
7
INPUT
R1 1MΩ
2
I_{IN}
CR1
CR2
8
ICH8500
6
OUTPUT
$V_o = I_{IN} \times 10^{12}\Omega$
$= -1$ VOLT/pA
5
INTERNAL DIODES
3
20kΩ
1
4
−15 V

Fig. 20-2

Circuit Notes

A very sensitive pico ammeter (−1 V/pA) employs the amplifier in the inverting or current summing mode. Care must be taken to eliminate stray currents from flowing into the current summing mode. It takes approximately 5 for the circuit to stabilize to within 1% of its final output voltage after a step function of input current has been applied. The internal diodes CR1 and CR2 together with external resistor R1 to protect the input stage of the amplifier from voltage transients.

NANO AMMETER


R_f
I_{IN}
+
−
+1.5V
2
7
1N914
D1
1N914
D2
LM4250
6
V_O
3
5
4
8
I_O
R1
5K
1
R2
100K
R_i
C1
0.1 μF
R_s
10M
M
−
+
−1.5V
+1.5V
B1
1.5V
POWER SUPPLY
B2
1.5V
−1.5V

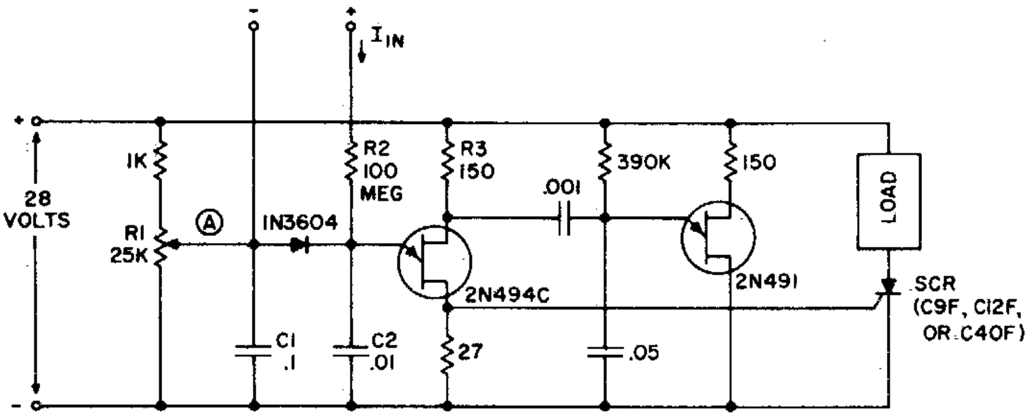
Resistance Values for DC Nano-and Micro Ammeter

I FULL SCALE	R_f (Ω)	R_f' (Ω)
100 nA	1.5M	1.5M
500 nA	300k	300k
1 μA	300k	0
5 μA	60k	0
10 μA	30k	0
50 μA	6k	0
100 μA	3k	0

The complete meter amplifier is a differential current-to-voltage converter with input protection, zeroing and full scale adjust provisions, and input resistor balancing for minimum offset voltage.

Fig. 20-3

NANOAMPERE SENSING CIRCUIT WITH 100 MEGOHM INPUT IMPEDANCE



-
+
I_{IN}
+
28 VOLTS
-
IK
RI 25K
(A)
IN3604
R2 100 MEG
R3 150
.001
390K
150
LOAD
2N494C
2N491
SCR (C9F, CI2F, OR C40F)
CI .I
C2 .01
27
.05

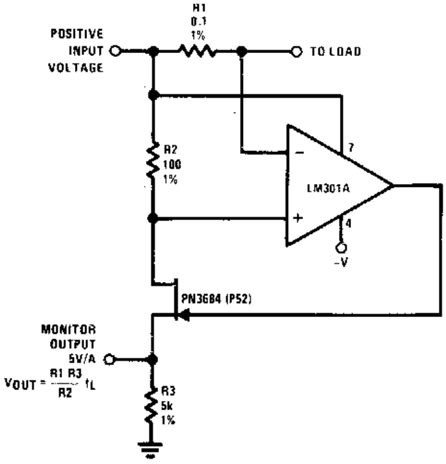
NANOAMPERE SENSING CIRCUIT WITH
100 MEGOHM INPUT IMPEDANCE

Fig. 20-4

Circuit Notes

The circuit may be used as a sensitive current detector or as a voltage detector having high input impedance. R1 is set so that the voltage at point (A) is ½ to ¾ volts below the level that fires the 2N494C. A small input current (Iin) of only 40 nanoamperes will charge C2 and raise the voltage at the emitter to the firing level. When the 2N494C fires, both capacitors, C1 and C2, are discharged through the 27 ohm resistor, which generates a positive pulse with sufficient amplitude to trigger a controlled rectifier (SCR), or other pulse sensitive circuitry.

CURRENT MONITOR



POSITIVE INPUT VOLTAGE
R1 0.1 1%
TO LOAD
R2 100 1%
LM301A
-
+
7
4
-V
PN3684 (P52)
MONITOR OUTPUT 5V/A
$V_{OUT} = \frac{R1\ R3}{R2} I_L$
R3 5k 1%

Circuit Notes

R1 senses current flow of a power supply. The JFET is used as a buffer because $I_D = I_S$; therefore the output monitor voltage accurately reflects the power supply current flow.

Fig. 20-5

21

Current Sources and Sinks

The sources of the following circuits are contained in the Sources section beginning on page 730. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

Current source
Precision Current Source
Precision 1 μA to 1 mA Current Sources
Precision Current Sink

CURRENT SOURCE

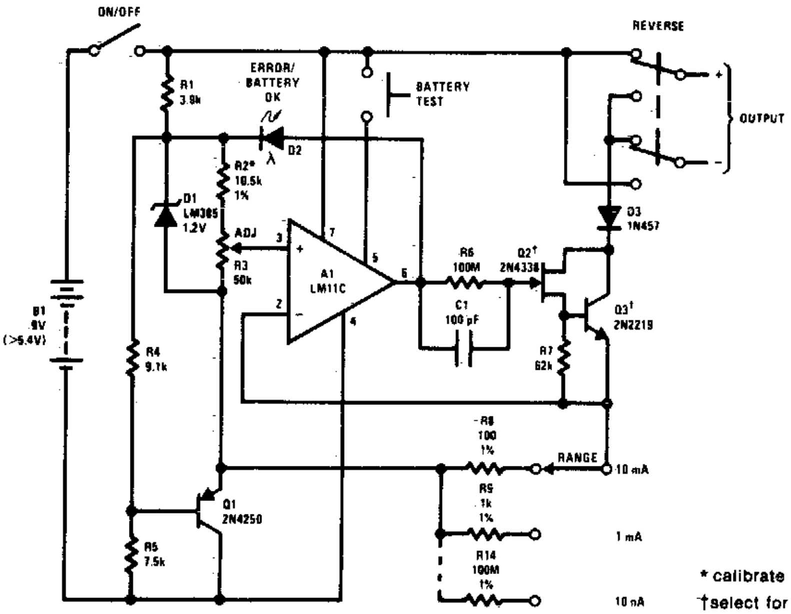

ON/OFF
REVERSE
R1
3.9k
ERROR/
BATTERY
OK
BATTERY
TEST
OUTPUT
+
-
D2
λ
R2*
10.5k
1%
D1
LM385
1.2V
ADJ
R3
50k
3
7
5
6
2
4
+
-
A1
LM11C
D3
1N457
R6
100M
Q2†
2N4338
C1
100 pF
Q3†
2N2219
R7
62k
B1
9V
(>5.4V)
R4
9.1k
R8
100
1%
RANGE
10 mA
R9
1k
1%
1 mA
Q1
2N4250
R5
7.5k
R14
100M
1%
10 nA

Fig. 21-1

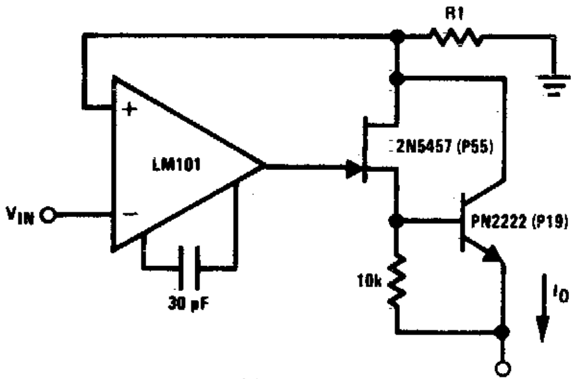
* calibrate range

†select for $I_{CBO} \leq 100$ pA

Circuit Notes

This precision current source has 10 μA to 10 mA ranges with output compliance of 30V to −5V. Output current is fully adjustable on each range with a calibrated, ten-turn potentiometer. Error light indicates saturation.

PRECISION CURRENT SOURCE


R1
+
-
LM101
2N5457 (P55)
V_{IN}
PN2222 (P19)
30 pF
10k
I_O
V^-

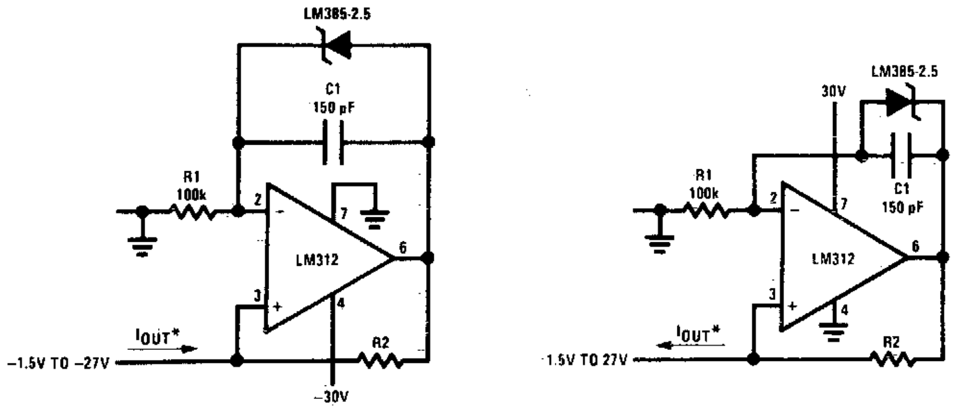
$$I_O = \frac{V_{IN}}{R1} \quad V_{IN} \leq 0V$$

Fig. 21-2

Circuit Notes

The 2N5457 and PN2222 bipolar serve as voltage isolation devices between the output and the current sensing resistor, R1. The LM101 provides a large amount of loop gain to assure that the circuit acts as a current source. For small values of current (<1 mA), the PN2222 and 10K resistor may be eliminated with the output appearing at the source of the 2N5457.

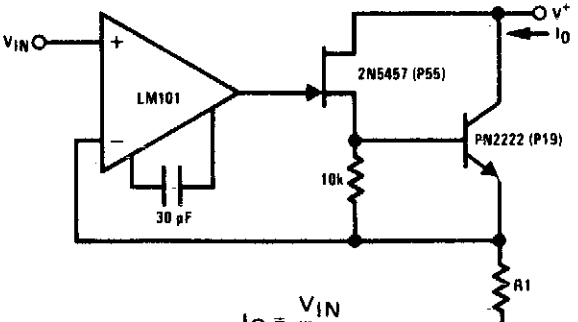
PRECISION 1 μA to 1 mA CURRENT SOURCES


LM385-2.5
C1
150 pF
R1
100k
2
7
6
3
4
LM312
I_{OUT}*
R2
−1.5V TO −27V
−30V
LM385-2.5
30V
R1
100k
C1
150 pF
2
7
6
3
4
LM312
I_{OUT}*
R2
1.5V TO 27V

$$* I_{OUT} = \frac{2.5V}{R2}$$

Fig. 21-3

PRECISION CURRENT SINK


V_{IN}
+
−
LM101
30 pF
2N5457 (P55)
10k
PN2222 (P19)
R1
V+
I_O

$$I_O = \frac{V_{IN}}{R1}$$

$$V_{IN} > 0V$$

Circuit Notes

The 2N5457 JFET and PN2222 bipolar have inherently high output impedance. Using R1 as a current sensing resistor to provide feedback to the LM101 op amp provides a large amount of loop gain for negative feedback to enhance the true current sink nature of this circuit. For small current values, the 10 K resistor and PN2222 may be eliminated if the source of the JFET is connected to R1.

Fig. 21-4

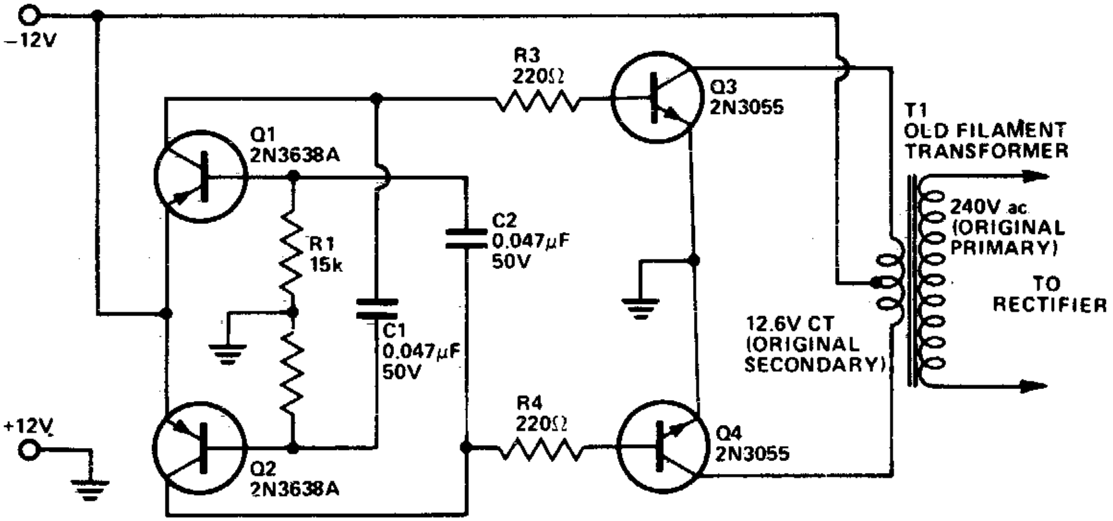
22

Dc/Dc and Dc/Ac Converters

The sources of the following circuits are contained in the Sources section beginning on page 730. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

Dc-to-Dc/Ac Inverter
Dc-to-Dc SMPS Using NE5561 Variable 18 V to 30 V Out at 0.2 A
Mini Power Inverter as High Voltage, Low Current Source
Regulated Dc-to-Dc Converter
400 V, 60 W Push-Pull Dc/Dc Converter
Dc/Dc Regulating Converter
Flyback Converter

DC-TO-DC/AC INVERTER



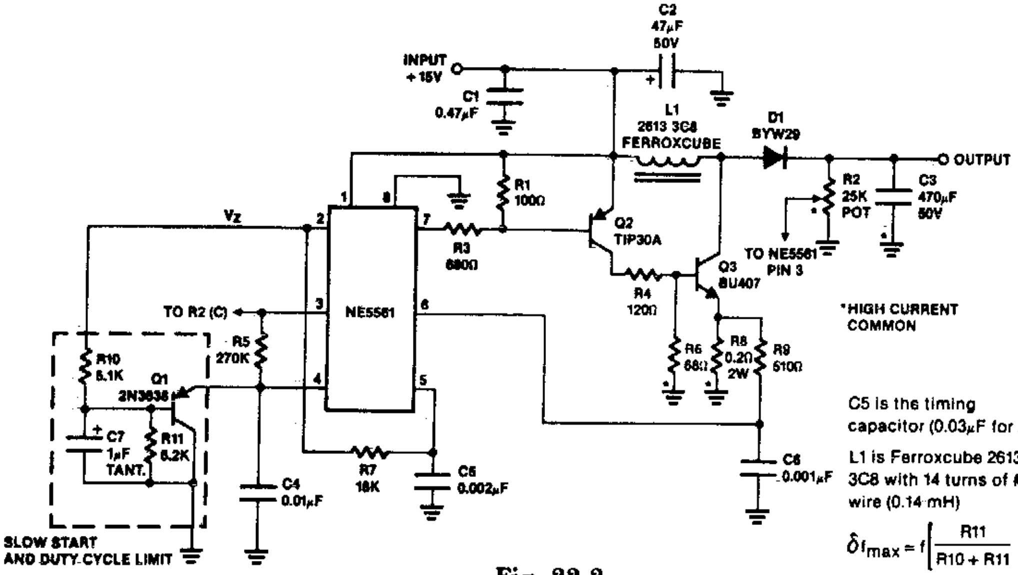
−12V
+12V
R3
220Ω
Q3
2N3055
T1
OLD FILAMENT TRANSFORMER
Q1
2N3638A
C2
0.047µF
50V
R1
15k
240V ac
(ORIGINAL PRIMARY)
TO RECTIFIER
C1
0.047µF
50V
12.6V CT
(ORIGINAL SECONDARY)
R4
220Ω
Q4
2N3055
Q2
2N3638A

Fig. 22-1

Circuit Notes

This inverter uses no special components such as the torodial transformer used in many inverters. Cost is kept low with the use of cheap, readily available components. Essentially, it is a power amplifier driven by an astable multivibrator. The frequency is around 1200 Hz which most 50/60 Hz power transformers handle well without too much loss. Increasing the value of capacitors C1 and C2 will lower the frequency if any trouble is experienced. However, rectifier filtering capacitors required are considerably smaller at the higher operating frequency. The two 2N3055 transistor should be mounted on an adequately sized heatsink. The transformer should be rated according to the amount of output power required allowing for conversion efficiency of approximately 60%.

DC-TO-DC SMPS USING NE5561 VARIABLE 18 V to 30 V OUT AT 0.2 A



C2
47μF
50V
INPUT
+15V
C1
0.47μF
L1
2613 3C8
FERROXCUBE
D1
BYW29
OUTPUT
R1
100Ω
R2
25K
POT
C3
470μF
50V
Vz
R3
680Ω
Q2
TIP30A
TO NE5561
PIN 3
Q3
BU407
R4
120Ω
NE5561
TO R2 (C)
*HIGH CURRENT
COMMON
R5
270K
R10
5.1K
Q1
2N3638
R6
68Ω
R8
0.2Ω
2W
R9
510Ω
C7
1μF
TANT.
R11
8.2K
R7
18K
C5
0.002μF
C4
0.01μF
C6
0.001μF
SLOW START
AND DUTY-CYCLE LIMIT
1
2
3
4
5
6
7
8

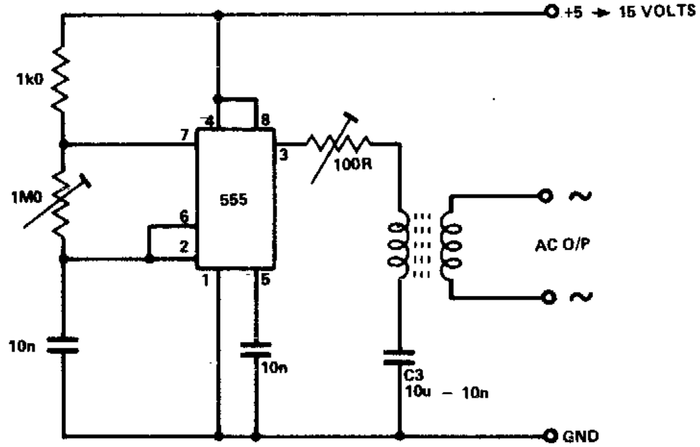
C5 is the timing capacitor (0.03μF for 22 kHz)

L1 is Ferroxcube 2613-3C8 with 14 turns of #18 enameled wire (0.14 mH)

$$\delta f_{max} = f\left[\frac{R11}{R10 + R11}(V_Z) + V_{BE}\right]$$

Fig. 22-2

MINI POWER INVERTER AS HIGH VOLTAGE, LOW CURRENT SOURCE



+5 → 15 VOLTS
1k0
1M0
10n
555
4
8
7
3
6
2
1
5
100R
AC O/P
~
~
10n
C3
10u – 10n
GND

Fig. 22-3

Circuit Notes

The circuit is capable of providing power for portable Geiger counters, dosimeter chargers, high resistance meters, etc. The 555 timer IC is used in its multivibrator mode, the frequency adjusted to optimize the transformer characteristics. When the output of the IC is high, current flows through the limiting resistor, the primary coil to charge C3. When the output is low, the current is reversed. With a suitable choice of frequency and C3, a good symmetric output is sustained.

REGULATED DC-TO-DC CONVERTER

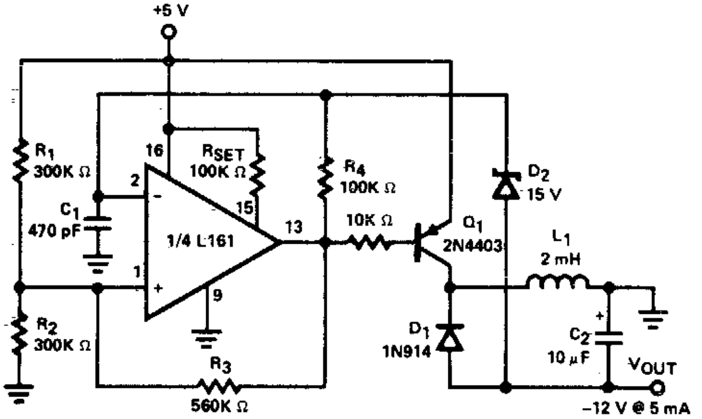
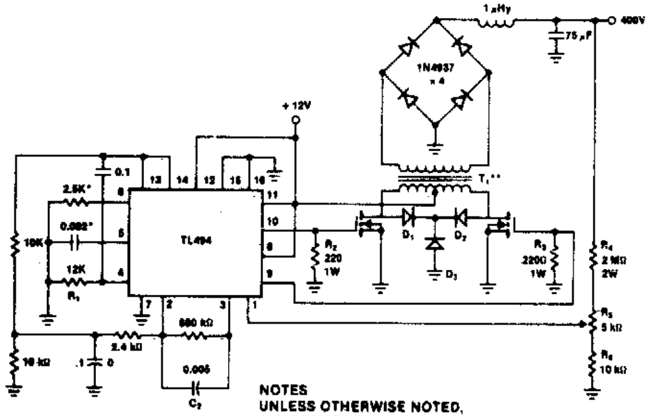

+5 V
R1
300K Ω
16
RSET
100K Ω
2
15
C1
470 pF
1/4 L161
13
R4
100K Ω
10K Ω
Q1
2N4403
D2
15 V
L1
2 mH
1
9
R2
300K Ω
R3
560K Ω
D1
1N914
C2
10 μF
VOUT
-12 V @ 5 mA

Fig. 22-4

Circuit Notes

Low power dc to dc converter obtained by adding a flyback circuit to a square wave oscillator. Operating frequency is 20 kHz to minimize the size of L1 and C2. Regulation is achieved by zener diode D2. Maximum current available before the converter drops out of regulation is 5.5 mA.

400 V, 60 W PUSH-PULL DC/DC CONVERTER


1 μHy
400V
75 μF
1N4937
× 4
+12V
T1**
0.1
2.5K*
0.002*
10K
12K
R1
TL494
13
14
12
15
16
11
10
8
9
6
5
4
7
2
3
1
D1
D2
D3
R2
220
1W
R3
2200
1W
R4
2 MΩ
2W
R5
5 kΩ
R6
10 kΩ
680 kΩ
2.4 kΩ
10 kΩ
.1
0.005
C2

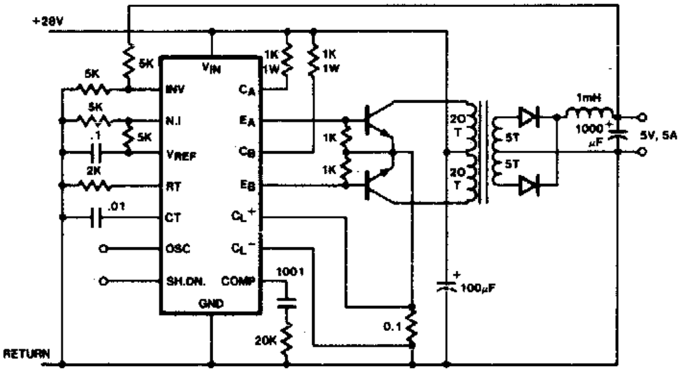
NOTES
UNLESS OTHERWISE NOTED,
ALL RESISTORS 5%, ¼ W
ALL CAPACITOR VALUES IN MICROFARADS, 25V
Q_1 & Q_2: VN64GA ON HEAT SINK
D_1 & D_2: 1N4934
D_3: 33V, 3W ZENER
T_1: PRI: 12T, CT, NO 18 AWG
SEC: 275T, NO 24 AWG
CORE: IND GEN 8231-1

Fig. 22-5

Circuit Notes

The TL494 switching regulator governs the operating frequency and regulates output voltage. Switching frequency approximately 100 kHz for the values shown. Output regulation is typically 1.25% from no-load to full 60 W.

DC/DC REGULATING CONVERTER



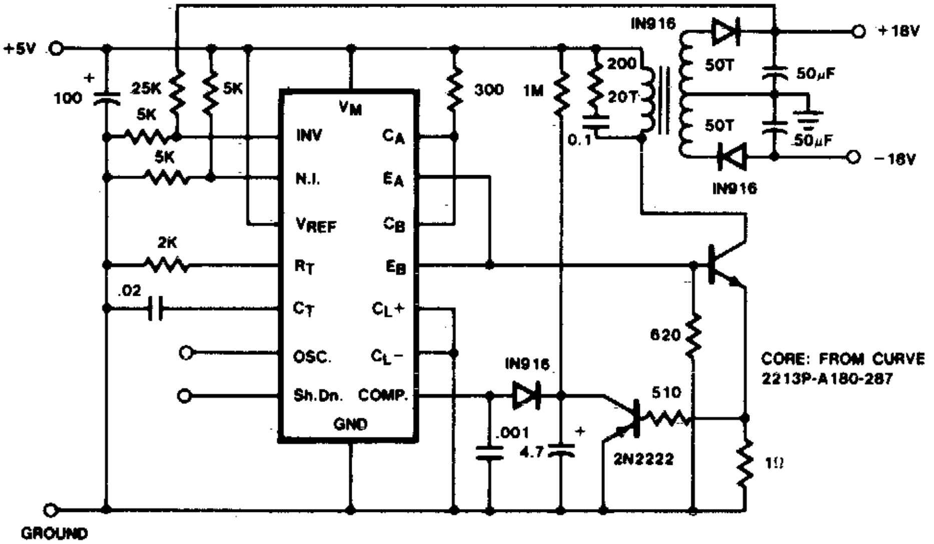
+28V
5K
5K
5K
.1
5K
2K
.01
RETURN
VIN
INV
N.I
VREF
RT
CT
OSC
SH.DN.
GND
CA
EA
CB
EB
CL+
CL−
COMP
1K 1W
1K 1W
1K
1K
1001
20K
0.1
20 T
20 T
5T
5T
1mH
1000 µF
5V, 5A
100µF

Fig. 22-6

Circuit Notes

Push-pull outputs are used in this transformer-coupled dc-dc regulating converter. Note that the oscillator must be set at twice the desired output frequency as the SG1524's internal flip-flop divides the frequency by 2 as it switches the PWM signal from one output to the other. Current limiting is done here in the primary so that the pulse width will be reduced should transformer saturation occur.

FLYBACK CONVERTER



+5V
100
25K
5K
5K
5K
2K
.02
GROUND
VM
INV
N.I.
VREF
RT
CT
OSC.
Sh.Dn.
GND
CA
EA
CB
EB
CL+
CL−
COMP.
300
1M
200
0.1
20T
50T
50T
IN916
IN916
50µF
50µF
+18V
−18V
620
IN916
510
.001
4.7
2N2222
1Ω
CORE: FROM CURVE
2213P-A180-287

Fig. 22-7

Circuit Notes

A low-current flyback converter is used here to generate ±15 volts at 20 mA from a +5 volt regulated line. The reference generator in the SG1524 is unused with the input voltage providing the reference. Current limiting in a flyback converter is difficult and is accomplished here by sensing current in the primary line and resetting a soft-start circuit.

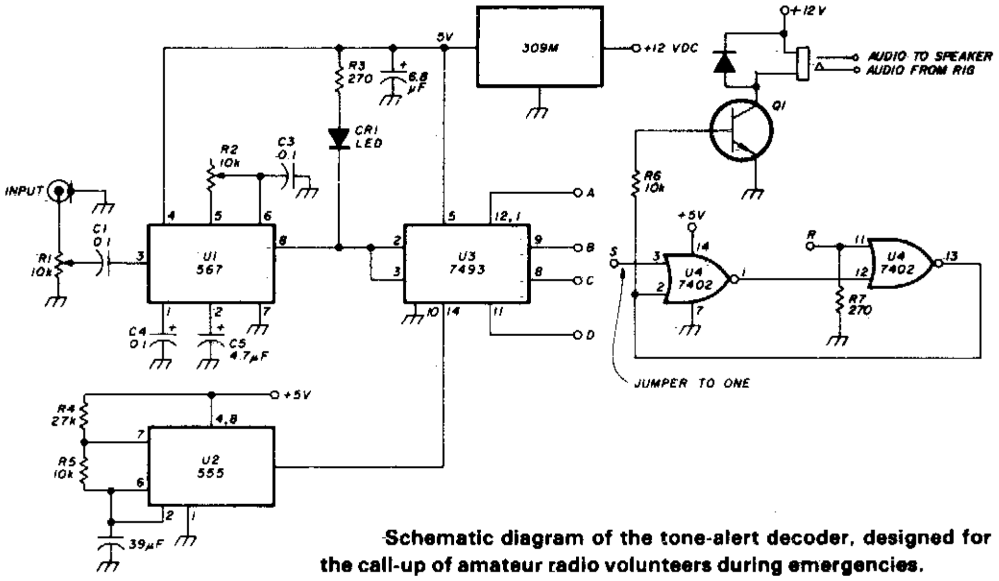
23

Decoders

The sources of the following circuits are contained in the Sources section beginning on page 730. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

Tone Alert Decoder
Tone Decoder with Relay Output
SCA Decoder
10.8 MHz FSK Decoder
24% Bandwidth Tone Decoder
Dual-Tone Decoder

TONE-ALERT DECODER


309M
5V
+12 VDC
+12 V
AUDIO TO SPEAKER
AUDIO FROM RIG
Q1
R3 270
6.8 µF
CR1 LED
R2 10k
C3 0.1
INPUT
C1 0.1
R1 10k
U1 567
U3 7493
A
B
C
D
R6 10k
+5V
S
R
U4 7402
U4 7402
R7 270
JUMPER TO ONE
C4 0.1
C5 4.7µF
R4 27k
R5 10k
+5V
U2 555
39µF

Schematic diagram of the tone-alert decoder, designed for the call-up of amateur radio volunteers during emergencies.

Fig. 23-1

Circuit Notes

PLL (U1) is set with R2 to desired tone frequency. LED lights to indicate lock-up of PLL. Reduce signal level (R1) and readjust R2 to assure lock-up. Delay is selected from counter U3 output. Circuits latches (turns on Q1 to allow audio to speaker) when proper frequency/duration signal is received. To reset latch, a positive voltage must be applied briefly to the R input of U4.

TONE DECODER WITH RELAY OUTPUT

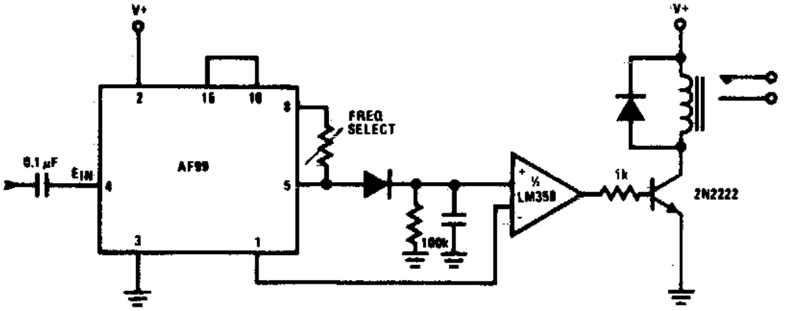

V+
V+
0.1 µF
E_{IN}
AF99
FREQ SELECT
100k
½ LM358
1k
2N2222

Fig. 23-2

SCA (BACKGROUND MUSIC) DECODER

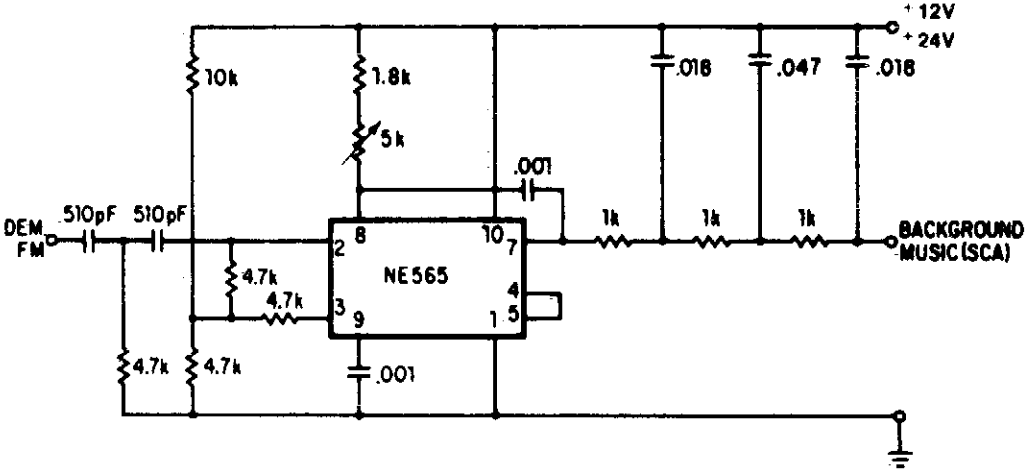

+12V
+24V
10k
1.8k
5k
.018
.047
.018
.001
DEM FM
510pF
510pF
8
10
2
7
NE565
4.7k
4.7k
4
3
9
1
5
1k
1k
1k
BACKGROUND MUSIC (SCA)
4.7k
4.7k
.001

Fig. 23-3

Circuit Notes

A resistive voltage divider is used to establish a bias voltage for the input (pins 2 and 3). The demodulated (multiplex) FM signal is fed to the input through a two-stage high-pass filter, both to effect capacitive coupling and to attenuate the strong signal of the regular channel. A total signal amplitude, between 80 mV and 300 mV, is required at the input. Its source should have an impedance of less than 10,000 ohms. The Phase Locked Loop is tuned to 67 kHz with a 5000 ohm potentiometer; only approximate tuning is required, since the loop will seek the signal. The demodulated output (pin 7) passes through a three-stage low-pass filter to provide de-emphasis and attenuate the high-frequency noise which often accompanies SCA transmission. The demodulated output signal is in the order of 50mV and the frequency response extends to 7 kHz.

10.8 MHz FSK DECODER

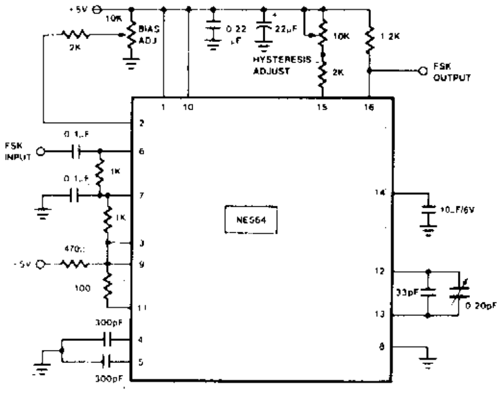
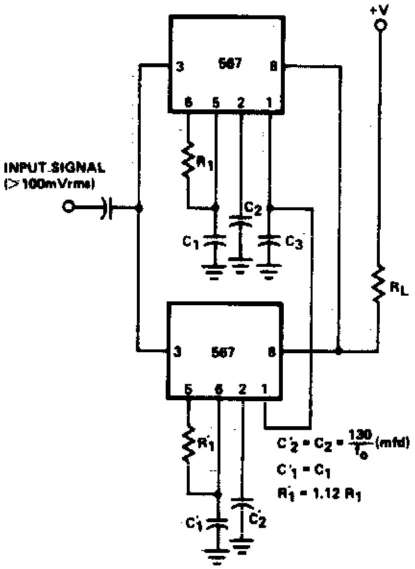

+5V
10K
BIAS ADJ
2K
0.22 μF
22μF
10K
1.2K
HYSTERESIS ADJUST
2K
FSK OUTPUT
1
10
15
16
2
0.1μF
FSK INPUT
6
1K
0.1μF
7
14
10μF/6V
NE564
1K
3
470Ω
-5V
9
100
12
33pF
0-20pF
11
13
300pF
4
5
300pF
8

Fig. 23-4

24% BANDWIDTH TONE DECODER


+V
567
3
8
6
5
2
1
INPUT SIGNAL
(>100mVrms)
R1
C1
C2
C3
RL
567
3
8
5
6
2
1
R'1
C'1
C'2

$C'_2 = C_2 = \frac{130}{f_o}$ (mfd)

$C'_1 = C_1$

$R'_1 = 1.12\,R_1$

Fig. 23-5

DUAL-TONE DECODER

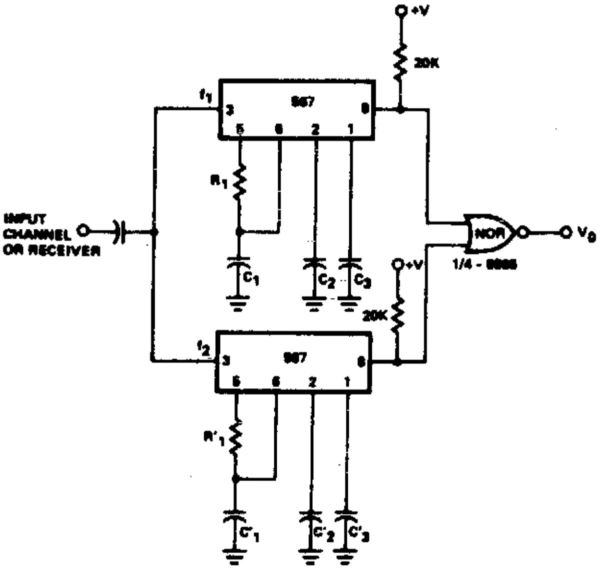

+V
20K
f1
3
567
8
5
6
2
1
R1
INPUT
CHANNEL
OR RECEIVER
C1
C2
C3
NOR
Vo
+V
1/4 -
20K
f2
3
567
8
5
6
2
1
R'1
C'1
C'2
C'3

Fig. 23-6

1. Resistor and capacitor-values chosen for desired frequencies and bandwidth.
2. If C_3 is made large so as to delay turn-on of the top 567, decoding of sequential (f_1 f_2) tones is possible.

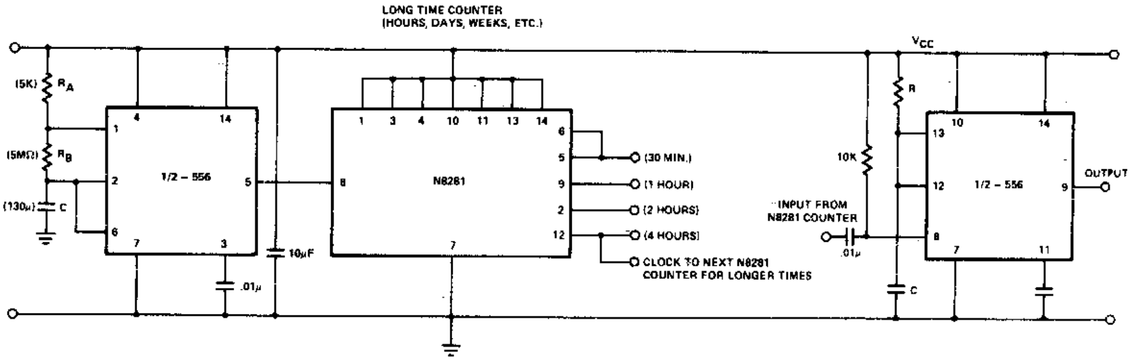
24

Delays

The sources of the following circuits are contained in the Sources section beginning on page 730. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

Long Time Delay
Time Delay Generator
Door Chimes Delay
Time Delay Generator
Long Delay Timer Using PUT
Ultra-Precise Long Time Delay Relay
Long Duration Time Delay
Simple Time Delay Using Two SCRs

LONG TIME DELAY


LONG TIME COUNTER
(HOURS, DAYS, WEEKS, ETC.)
(5K)
R_A
(5MΩ)
R_B
(130μ)
C
1/2 – 556
4
14
1
2
5
6
7
3
.01μ
10μF
N8281
1
3
4
10
11
13
14
6
5
9
2
12
8
7
(30 MIN.)
(1 HOUR)
(2 HOURS)
(4 HOURS)
CLOCK TO NEXT N8281
COUNTER FOR LONGER TIMES
V_{CC}
R
10K
INPUT FROM
N8281 COUNTER
.01μ
C
1/2 – 556
10
14
13
12
8
7
11
9
OUTPUT

TOTAL PERIOD = .693 $(R_A + 2R_B)$ C.
(EXAMPLE VALUES SET FOR APPROX. 15 MIN.)

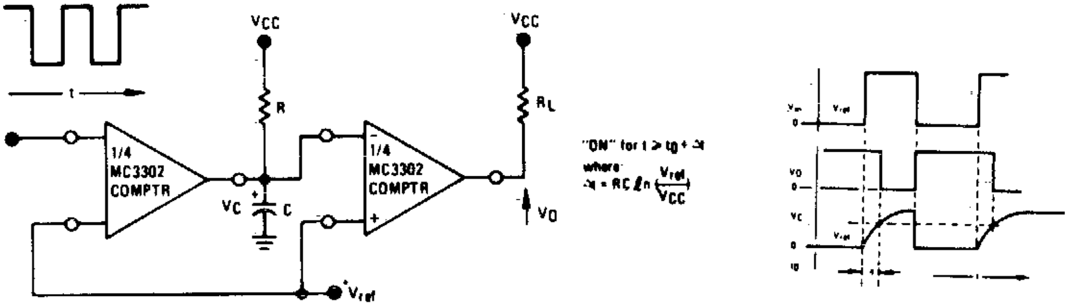
OUTPUT PULSE WIDTH, T = 1.1RC

Fig. 24-1

Circuit Notes

In the 556 timer, the timing is a function of the charging rate of the external capacitor. For long time delays, expensive capacitors with extremely low leakage are required. The practicality of the components involved limits the time between pulses to something in the neighborhood of 10 minutes. To achieve longer time periods, both halves of a dual timer may be connected in tandem with a "Divide-by"' network in between the first timer section operates in an oscillatory mode with a period of 1/fo. This signal is then applied to a "Divide-by-N" network to give an output with the period of N/fo. This can then be used to trigger the second half of the 556. The total time delay is now a function of N and fo.

TIME DELAY GENERATOR


t
VCC
R
1/4
MC3302
COMPTR
VC
C
VCC
R_L
1/4
MC3302
COMPTR
VO
V_{ref}
V_{in}
V_{ref}
0
VO
0
VC
V_{ref}
0
t_0
t

"ON" for $t > t_0 + \Delta t$
where
$\Delta t = RC \ln\left(\frac{V_{ref}}{V_{CC}}\right)$

Fig. 24-2

DOOR CHIMES DELAY

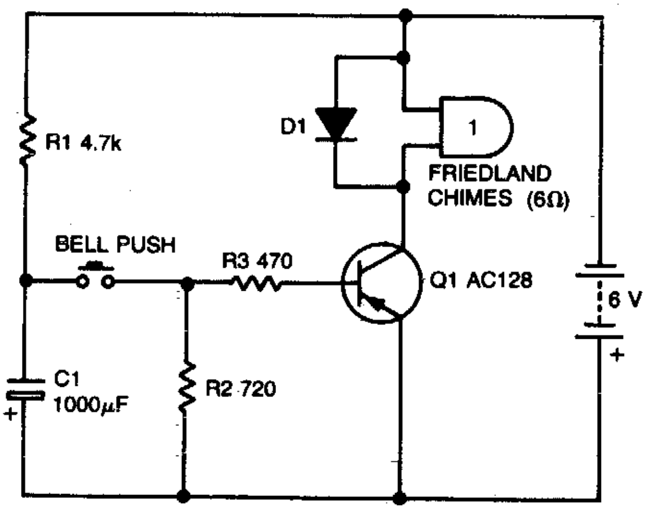

R1 4.7k
D1
1
FRIEDLAND
CHIMES (6Ω)
BELL PUSH
R3 470
Q1 AC128
6 V
+
C1
1000μF
+
R2 720

Fig. 24-3

Circuit Notes

With values shown, this simple circuit will permit one operation every 10 seconds or so. Capacitor C1 charges through R1 when the button is released. Making R1 larger will increase the delay.

TIME DELAY GENERATOR

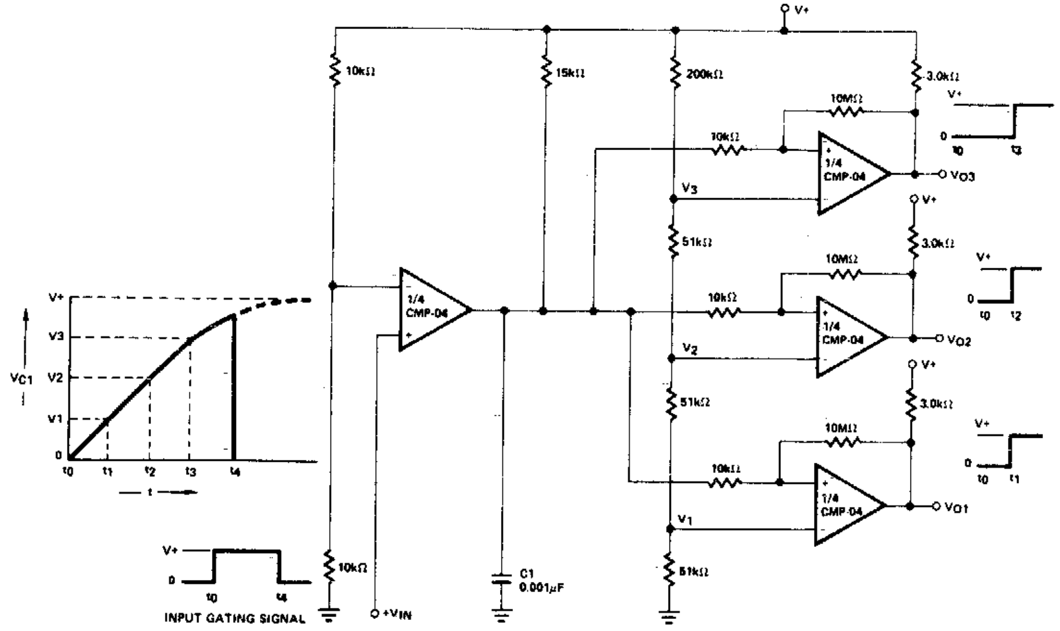

V+
10kΩ
15kΩ
200kΩ
3.0kΩ
10MΩ
10kΩ
1/4 CMP-04
VO3
V3
51kΩ
V+
3.0kΩ
10MΩ
10kΩ
1/4 CMP-04
VO2
V2
51kΩ
V+
3.0kΩ
10MΩ
10kΩ
1/4 CMP-04
VO1
V1
51kΩ
V+
0
t0
t3
V+
0
t0
t2
V+
0
t0
t1
1/4 CMP-04
C1
0.001μF
10kΩ
+VIN
V+
V3
V2
V1
0
VC1
t0
t1
t2
t3
t4
t
V+
0
t0
t4
INPUT GATING SIGNAL

Fig. 24-4

LONG DELAY TIMER USING PUT

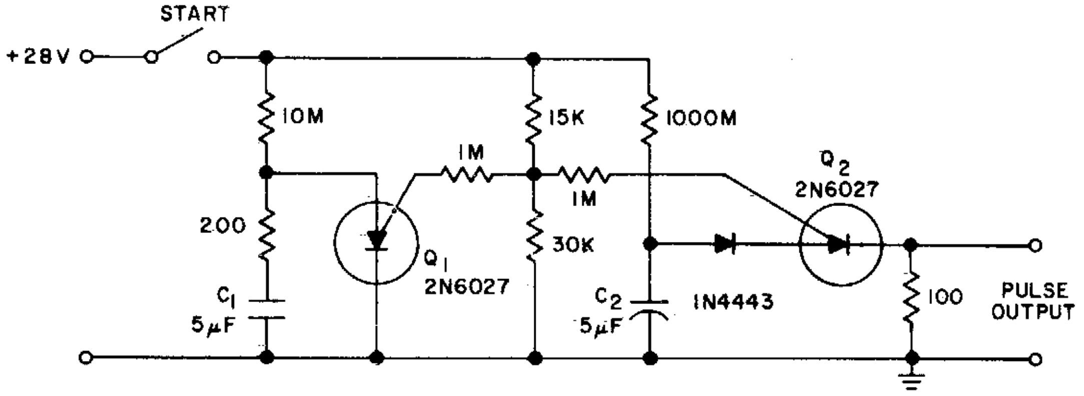

START
+28V
10M
15K
1000M
1M
1M
Q2
2N6027
200
Q1
2N6027
30K
C1
5μF
C2
5μF
1N4443
100
PULSE
OUTPUT

Fig. 24-5

Circuit Notes

The PUT is used as both a timing element and sampling oscillator. A low leakage film capacitor is required for C2 due to the low current supplied to it.

ULTRA-PRECISE LONG TIME DELAY RELAY

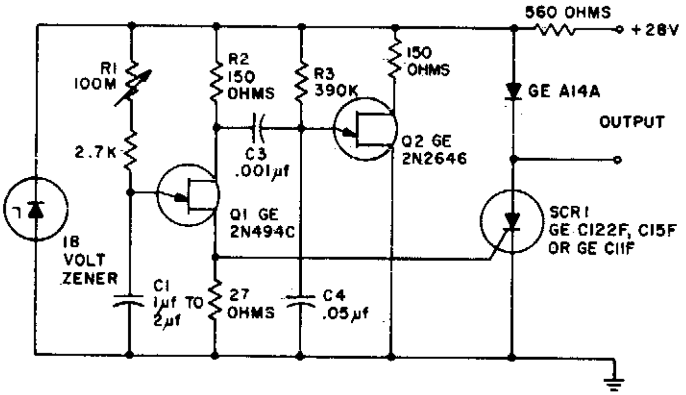

560 OHMS
+28V
R1
100M
R2
150
OHMS
R3
390K
150
OHMS
GE A14A
OUTPUT
2.7K
C3
.001μf
Q2 GE
2N2646
Q1 GE
2N494C
SCR1
GE C122F, C15F
OR GE C11F
18
VOLT
ZENER
C1
1μf TO
2μf
27
OHMS
C4
.05μf

Fig. 24-6

Circuit Notes

Predictable time delays from as low as 0.3 milliseconds to over 3 minutes are obtainable without resorting to a large value electrolytic-type timing capacitor. Instead, a stable low leakage paper or mylar capacitor is used and the peak point current of the timing UJT (Q1) is effectively reduced, so that a large value emitter resistor (R1) may be substituted.

LONG DURATION TIME DELAY

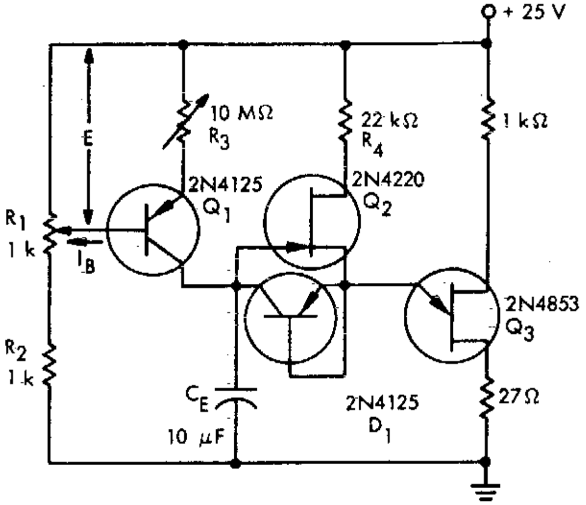

+ 25 V
E
10 MΩ
R_3
22 kΩ
R_4
1 kΩ
2N4125
Q_1
2N4220
Q_2
R_1
1 k
I_B
2N4853
Q_3
R_2
1 k
C_E
10 μF
2N4125
D_1
27 Ω

Fig. 24-7

SIMPLE TIME DELAY

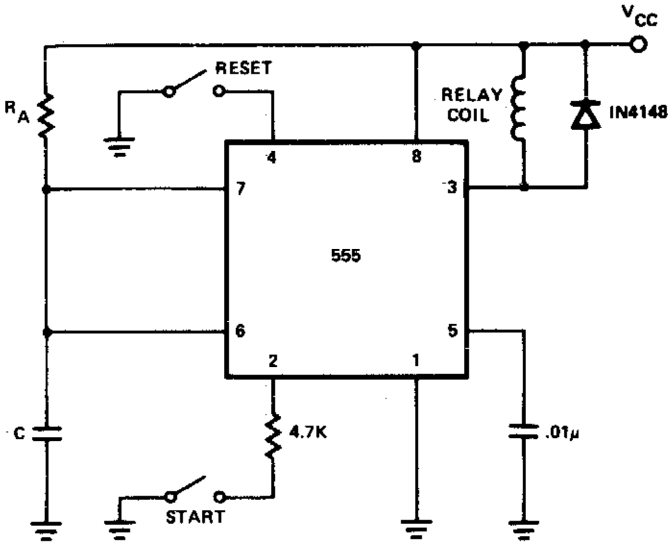

V_{CC}
RESET
R_A
RELAY
COIL
IN4148
4
8
7
3
555
6
5
2
1
C
4.7K
.01μ
START

Fig. 24-8

25

Detectors

The sources of the following circuits are contained in the Sources section beginning on page 730. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

Air-Motion Detector
Product Detector
Low Voltage Detector
Positive Peak Detector
Negative Peak Detector
Precision Peak Voltage Detector With Along Memory Time
Edge Detector
Ultra-Low Drift Peak Detector
Pulse Width Discriminator
True RMS Detector
Fast Half Wave Rectifier
Telemetry Demodulator
Full-Wave Rectifier and Averaging Filter
Double-Ended Limit Detector
Half-Wave Rectifier
Tone Detector
FM Tuner with a Single-Tuned Detector Coil
Missing Pulse Detector
High Speed Peak Detector
Detector for Magnetic Transducer
Double-Ended Limit Detector
FM Demodulator at 5 V
FM Demodulator at 12 V
Precision Full-Wave Rectifier
Negative Peak Detector
Level Detector with Hysteresis
Window Detector
Air Flow Detector
Positive Peak Detector

AIR-MOTION DETECTOR

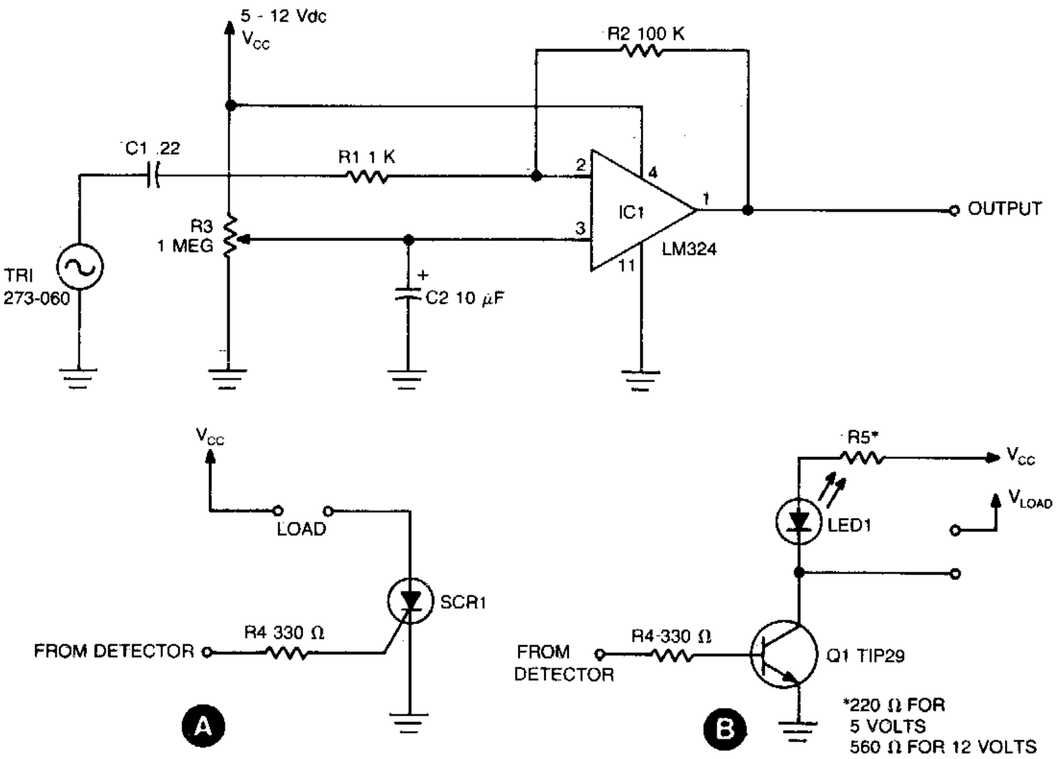

5 - 12 Vdc
V_{CC}
R2 100 K
C1 .22
R1 1 K
2
4
1
IC1
OUTPUT
R3
1 MEG
3
LM324
11
TRI
273-060
+
C2 10 μF
V_{CC}
LOAD
SCR1
R4 330 Ω
FROM DETECTOR
A
R5*
V_{CC}
V_{LOAD}
LED1
R4 330 Ω
FROM
DETECTOR
Q1 TIP29
*220 Ω FOR
5 VOLTS
560 Ω FOR 12 VOLTS
B

Fig. 25-1

Circuit Notes

Sensing circuit detects either steady or fluctuating air flows. The heart of the circuit is a Radio Shack piezo buzzer (P/N 273-060) and an LM324 quad op amp. (Red wire from the piezo element connects to capacitor C1, and the black wire to ground.) When a current of air hits the piezo element, a small signal is generated and is fed through C1 and R1 to the inverting input (pin 2) of one section of the LM324. That causes the output (pin 1) to go high. Resistor R3 adjusts sensitivity. The circuit can be made sensitive enough to detect the wave of a hand or the sensitivity can be set so low that blowing on the element hard will produce no output. Resistor R2 is used to adjust the level of the output voltage at pin 1. The detector circuit can be used in various control applications. For example, an SCR can be used to control 117-volt AC loads as shown in A. Also, an NPN transistor, such as a TIP29, can be used to control loads as shown in B.

PRODUCT DETECTOR

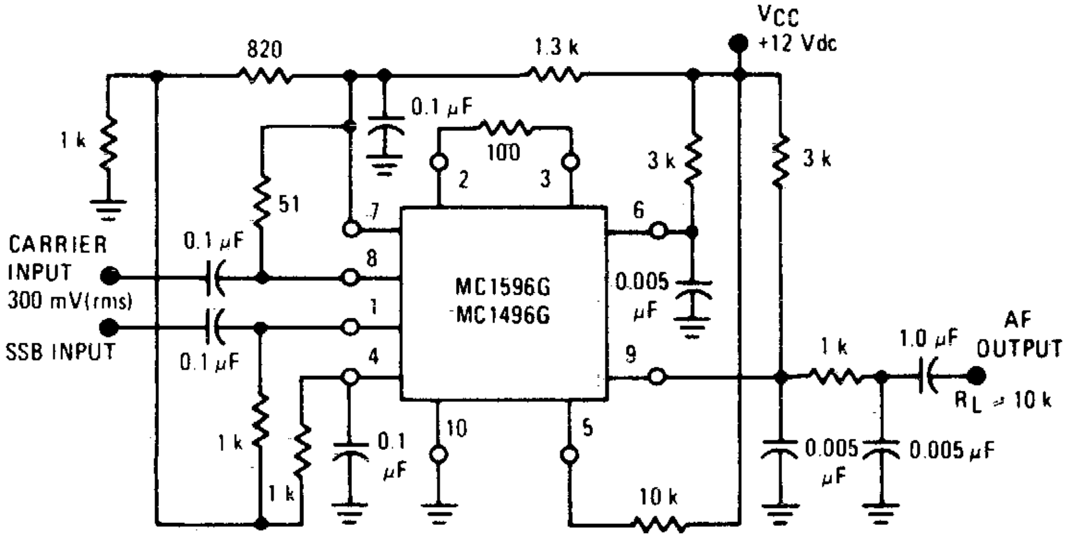

VCC
+12 Vdc
820
1.3 k
1 k
0.1 µF
100
2
3
3 k
3 k
51
7
6
CARRIER
INPUT
300 mV(rms)
0.1 µF
8
MC1596G
MC1496G
0.005
µF
1
AF
OUTPUT
SSB INPUT
0.1 µF
4
9
1 k
1.0 µF
RL ≥ 10 k
1 k
0.1
µF
10
5
0.005
µF
0.005 µF
1 k
10 k

Fig. 25-2

Circuit Notes

The MC1596/MC1496 makes an excellent SSB product detector. This product detector has a sensitivity of 3.0 microvolts and a dynamic range of 90 dB when operating at an intermediate frequency of 9 MHz. The detector is broadband for the entire high frequency range. For operation at very low intermediate frequencies down to 50 kHz the 0.1 µF capacitors on pins 7 and 8 should be increased to 1.0 µF. Also, the output filter at pin 9 can be tailored to a specific intermediate frequency and audio amplifier input impedance. The emitter resistance between pins 2 and 3 may be increased or decreased to adjust circuit gain, sensitivity, and dynamic range. This circuit may also be used as an AM detector by introducing carrier signal at the carrier input and an AM signal at the SSB input. The carrier signal may be derived from the intermediate frequency signal or generated locally. The carrier signal may be introduced with or without modulation, provided its level is sufficiently high to saturate the upper quad differential amplifier. If the carrier signal is modulated, a 300 mV (rms) input level is recommended.

LOW VOLTAGE DETECTOR

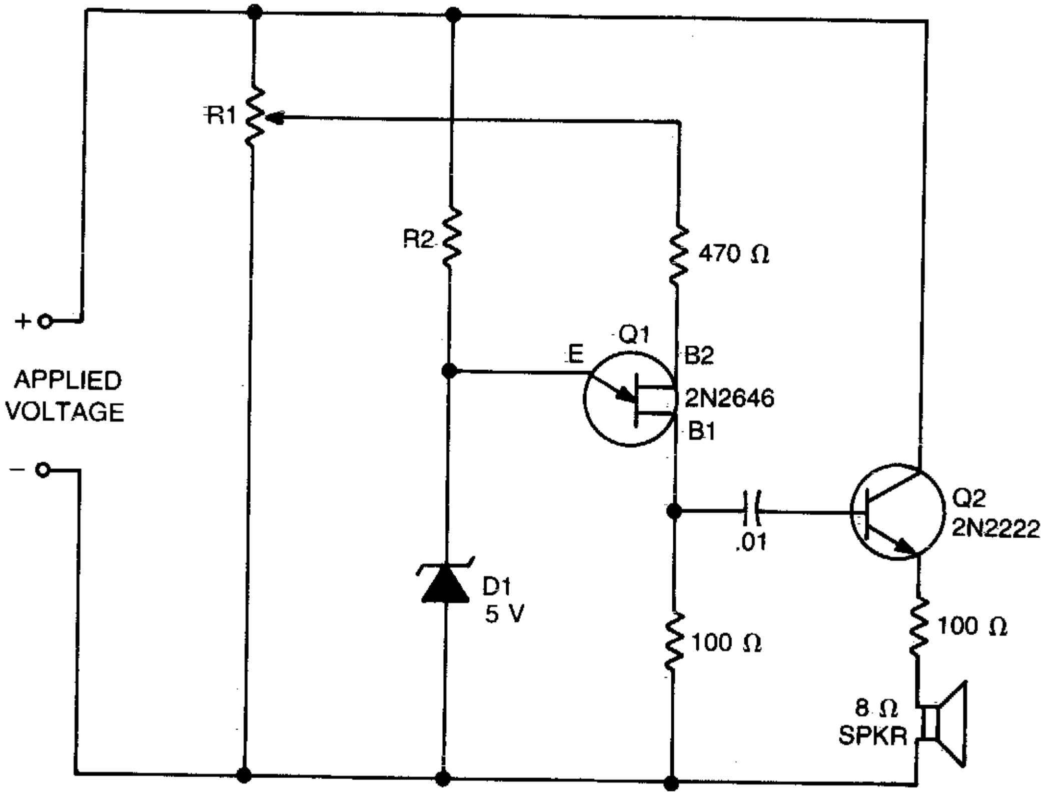
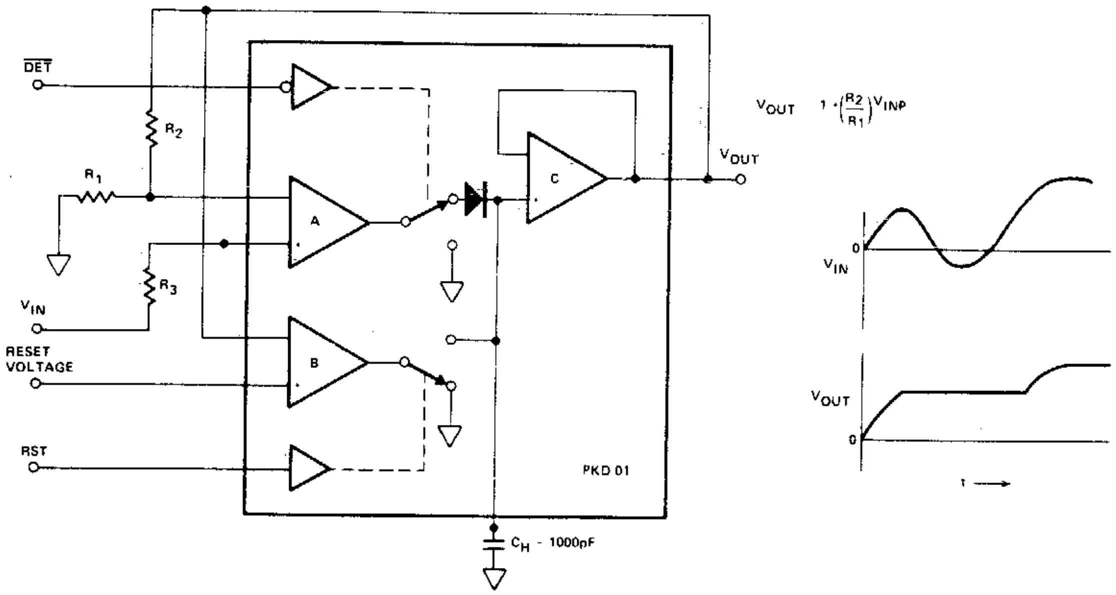


Fig. 25-3

Circuit Notes

The values of R1, R2, and D1 are selected for the voltage applied. Using a 12-volt battery, R1 = 10 K, R2 = 5.6 K and D1 is a 5-volt zener diode, or a string of forward-biased silicon rectifiers equaling about 5 volts. Transistor Q1 is a general-purpose UJT (Unijunction Transistor), and Q2 is any small-signal or switching NPN transistor. When detector is connected across the battery terminals, it draws little current and does not interfere with other devices powered by the battery. If voltage drops below the trip voltage selected with the R1 setting, the speaker beeps a warning. The frequency of the beeps is determined by the amount of undervoltage. If other voltages are being monitored, select R1 so that it draws only 1 mA or 2 mA. Zener diode D1 is about one-half of the desired trip voltage, and R2 is selected to bias it about 1 mA.

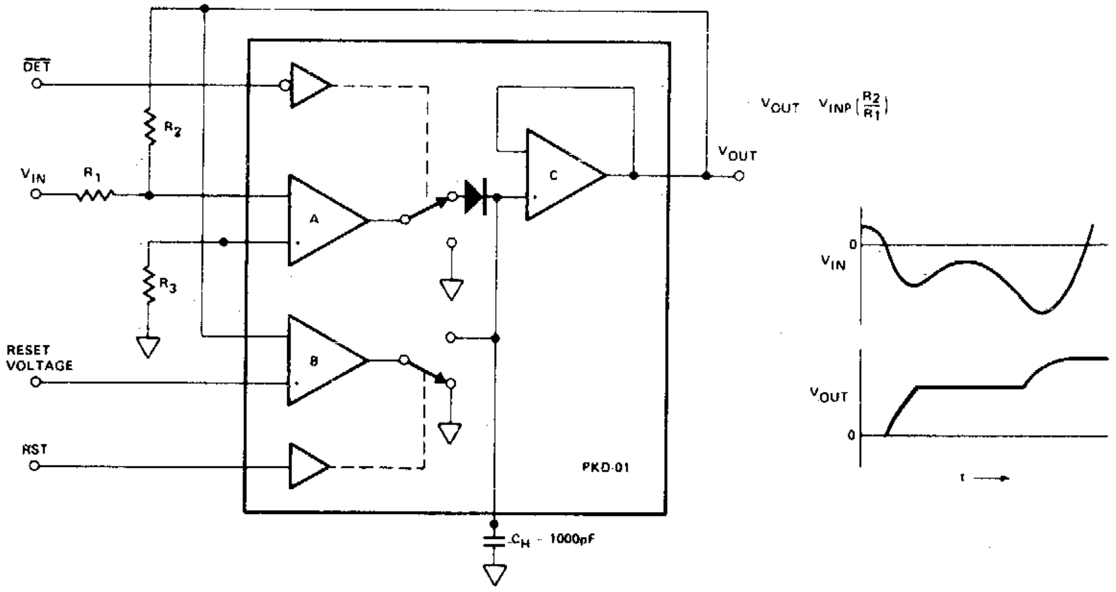
POSITIVE PEAK DETECTOR



$\overline{DET}$
R_2
R_1
V_{IN}
R_3
RESET VOLTAGE
RST
A
B
C
PKD 01
C_H - 1000pF
V_{OUT}
V_{OUT} $1 + \left(\frac{R2}{R1}\right) V_{INP}$
0
V_{IN}
V_{OUT}
0
t ⟶

Fig. 25-4

NEGATIVE PEAK DETECTOR



$\overline{DET}$
R_2
V_{IN}
R_1
R_3
RESET VOLTAGE
RST
A
B
C
PKD-01
C_H - 1000pF
V_{OUT}
V_{OUT} $V_{INP} \left(\frac{R2}{R1}\right)$
0
V_{IN}
V_{OUT}
0
t ⟶

Fig. 25-5

PRECISION PEAK VOLTAGE DETECTOR WITH A LONG MEMORY TIME

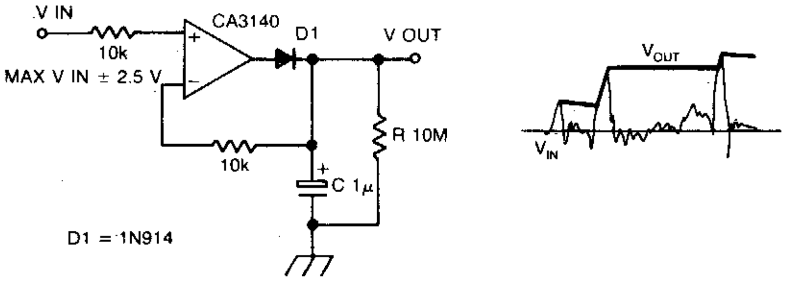
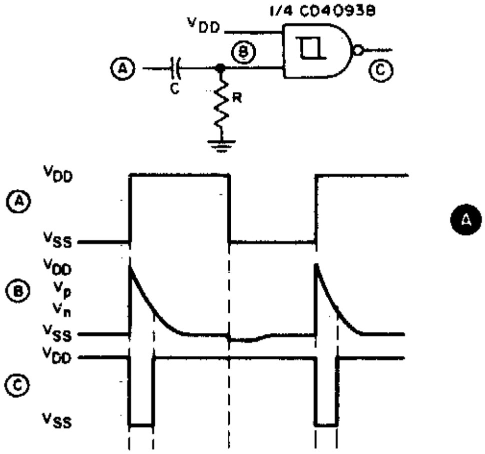

V IN
10k
MAX V IN ± 2.5 V
CA3140
D1
V OUT
R 10M
10k
C 1μ
D1 = 1N914
V_{OUT}
V_{IN}

Fig. 25-6

Circuit Notes

The circuit has negative feedback only for positive signals. The inverting input can only get some feedback when diode D1 is forward biased and only occurs when the input is positive. With a positive input signal, the output of the op amp rises until the inverting input signal reaches the same potential. In so doing, the capacitor C is also charged to this potential. When the input goes negative, the diode D1 becomes reverse biased, the voltage on the capacitor remains, being slowly discharged by the op amp input bias current of 10 pico amps. Thus the discharge of the capacitor is dominantly controlled by the resistor R, giving a time constant of 10 seconds. Thus, the circuit detects the most positive peak voltage and remembers it.

EDGE DETECTOR


1/4 CD4093B
V_{DD}
A
B
C
C
R
A
V_{DD}
V_{SS}
B
V_{DD}
V_p
V_n
V_{SS}
V_{DD}
C
V_{SS}
A

Circuit Notes

This circuit provides a short negative-going output pulse for every positive-going edge at the input. The input waveform is coupled to the input by capacitor C; the pulse length depends, as before, on R and C. If a negative going edge detector is required, the circuit in B should be used.

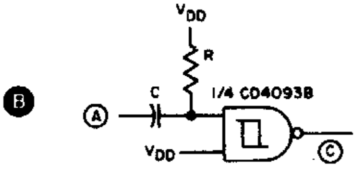
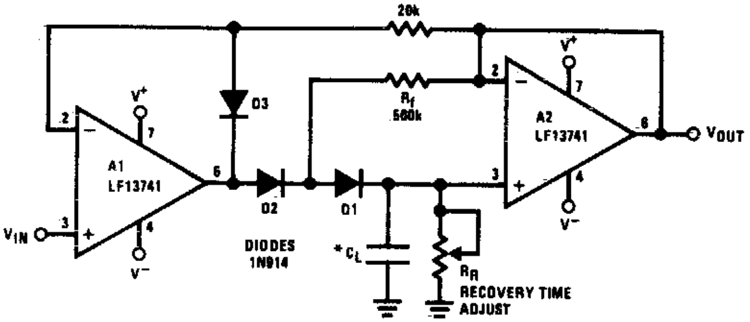

V_{DD}
R
B
C
1/4 CD4093B
A
V_{DD}
C

Fig. 25-7

ULTRA-LOW DRIFT PEAK DETECTOR

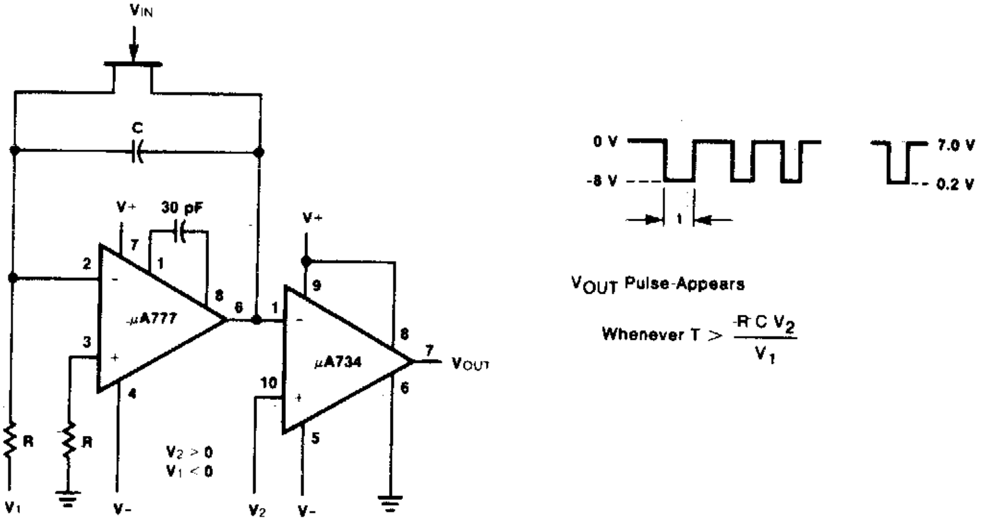

20k
V+
7
2
A1
LF13741
6
3
4
V−
V_IN
D3
D2
D1
R_f
560k
DIODES
1N914
*C_L
R_R
RECOVERY TIME
ADJUST
A2
LF13741
V+
V−
V_OUT

- By adding D1 and R_f, $V_{D1} = 0$ during hold mode. Leakage of D2 provided by feedback path through R_f.
- Leakage of circuit is I_B plus leakage of C_h.
- D3 clamps V_{OUT} A1 to $V_{IN} - V_{D3}$ to improve speed and to limit the reverse bias of D2.
- Maximum input frequency should be $<< 1/2\pi R_f C_{D2}$, where C_{D2} is the shunt capacitance of D2.

*Low leakage capacitor

Fig. 25-8

PULSE WIDTH DISCRIMINATOR

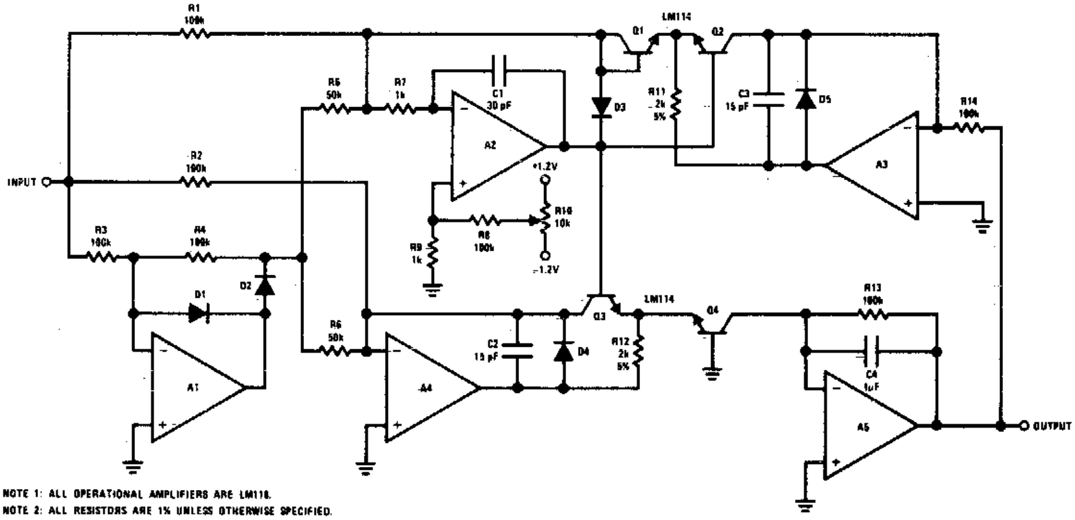

V_IN
C
V+
30 pF
7
1
8
6
2
3
4
μA777
V+
9
1
8
7
10
6
5
μA734
V_OUT
R
R
V₂ > 0
V₁ < 0
V₁
V−
V₂
V−
0 V
−8 V
T
7.0 V
0.2 V

V_{OUT} Pulse Appears

Whenever $T > \dfrac{-R\,C\,V_2}{V_1}$

Fig. 25-9

TRUE RMS DETECTOR


R1
100k
R2
100k
R3
100k
R4
100k
R5
50k
R6
50k
R7
1k
R8
100k
R9
1k
R10
10k
R11
2k
5%
R12
2k
5%
R13
100k
R14
100k
C1
30 pF
C2
15 pF
C3
15 pF
C4
1μF
D1
D2
D3
D4
D5
Q1
Q2
Q3
Q4
LM114
LM114
A1
A2
A3
A4
A5
+1.2V
−1.2V
INPUT
OUTPUT

NOTE 1: ALL OPERATIONAL AMPLIFIERS ARE LM118.
NOTE 2: ALL RESISTORS ARE 1% UNLESS OTHERWISE SPECIFIED.
NOTE 3: ALL DIODES ARE 1N914.
NOTE 4: SUPPLY VOLTAGE ±15V.

Fig. 25-10

Circuit Notes

The circuit will provide a dc output equal to the rms value of the input. Accuracy is typically 2% for a 20 V_{PP} input signal from 50 Hz to 100 kHz, although it's usable to about 500 kHz. The lower frequency is limited by the size of the filter capacitor. Since the input is dc coupled, it can provide the true rms equivalent of a dc and ac signal.

FAST HALF-WAVE RECTIFIER

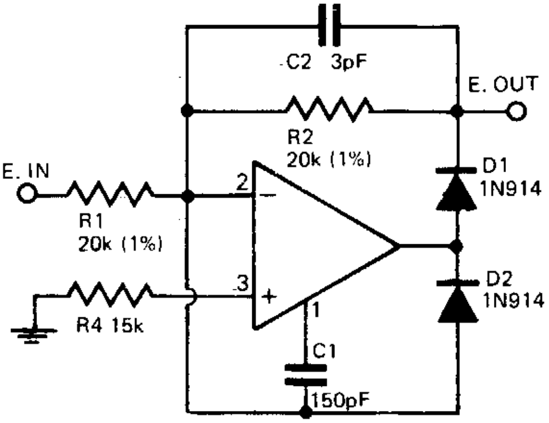
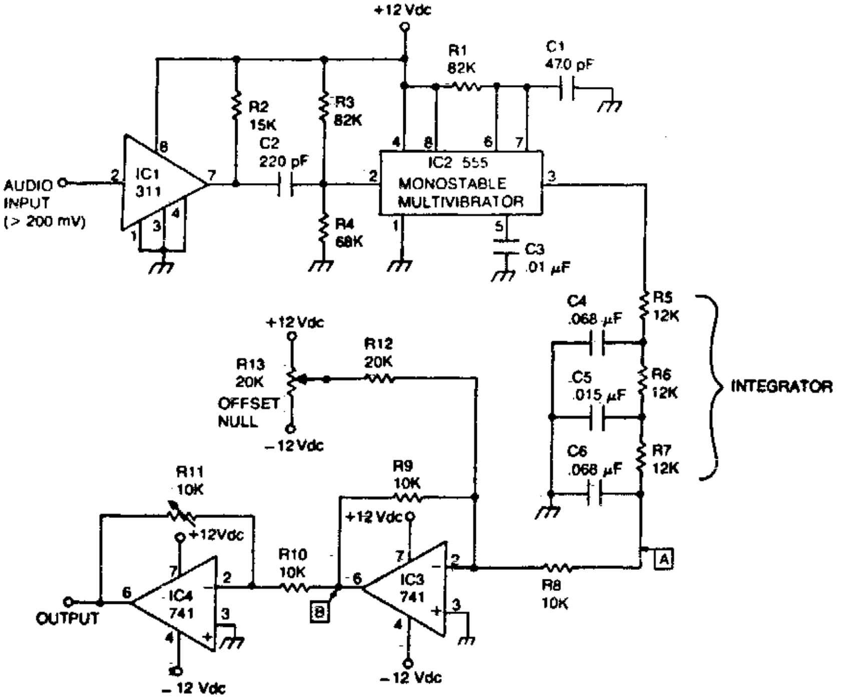

C2 3pF
E. OUT
R2
20k (1%)
D1
1N914
E. IN
2
R1
20k (1%)
3
D2
1N914
R4 15k
1
C1
150pF

Fig. 25-11

Circuit Notes

Precision half wave rectifier using an operational amplifier will have a rectification accuracy of 1% from dc to 100 kHz.

TELEMETRY DEMODULATOR



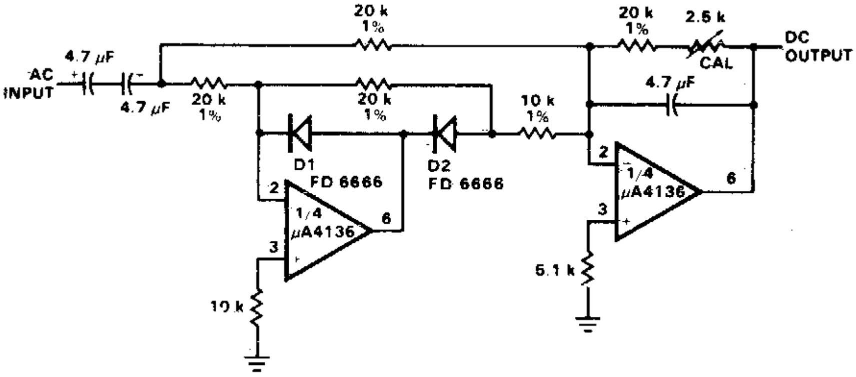
+12 Vdc
R1
82K
C1
470 pF
R2
15K
R3
82K
C2
220 pF
IC1
311
AUDIO
INPUT
(> 200 mV)
IC2 555
MONOSTABLE
MULTIVIBRATOR
R4
68K
C3
.01 μF
C4
.068 μF
R5
12K
C5
.015 μF
R6
12K
C6
.068 μF
R7
12K
INTEGRATOR
+12 Vdc
R13
20K
OFFSET
NULL
-12 Vdc
R12
20K
R11
10K
R9
10K
+12Vdc
+12 Vdc
R10
10K
IC4
741
IC3
741
OUTPUT
B
A
R8
10K
-12 Vdc
-12 Vdc

Fig. 25-12

Circuit Notes

The circuit recovers an FM audio signal that varies from less than 1 kHz to about 10 kHz.

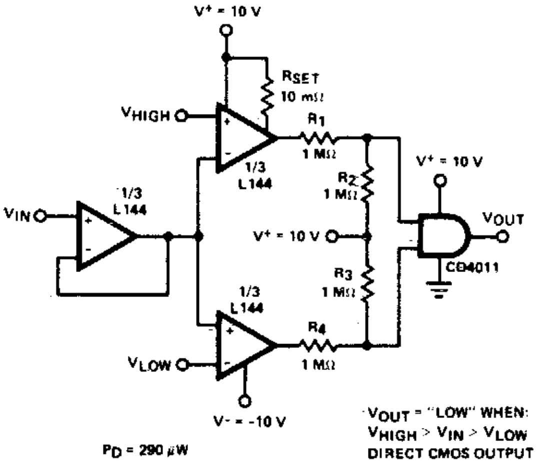
FULL-WAVE RECTIFIER AND AVERAGING FILTER



AC
INPUT
4.7 μF
4.7 μF
20 k
1%
20 k
1%
20 k
1%
D1
FD 6666
D2
FD 6666
10 k
1%
1/4
μA4136
10 k
20 k
1%
2.5 k
CAL
DC
OUTPUT
4.7 μF
1/4
μA4136
5.1 k

Fig. 25-13

DOUBLE-ENDED LIMIT DETECTOR

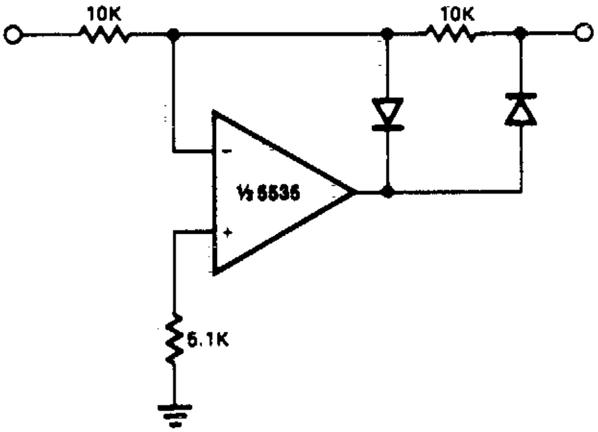

V+ = 10 V
RSET
10 mΩ
VHIGH
R1
1 MΩ
1/3
L144
V+ = 10 V
R2
1 MΩ
1/3
L144
VIN
VOUT
V+ = 10 V
CD4011
R3
1 MΩ
1/3
L144
R4
1 MΩ
VLOW
V- = -10 V
VOUT = "LOW" WHEN:
VHIGH > VIN > VLOW
DIRECT CMOS OUTPUT
PD = 290 μW

Circuit Notes

Detector uses three sections of an L144 and a CMOS NAND gate to make a very low power voltage monitor. The 1 MΩ resistors R1, R2, R3, and R4 translate the bipolar ±10 V swing of the op amps to a 0 to 10 V swing acceptable to the ground-referenced CMOS logic. The total power dissipation is 290 μW while in limit and 330 μW while out of limit.

Fig. 25-14

HALF-WAVE RECTIFIER


10K
10K
½ 5535
5.1K

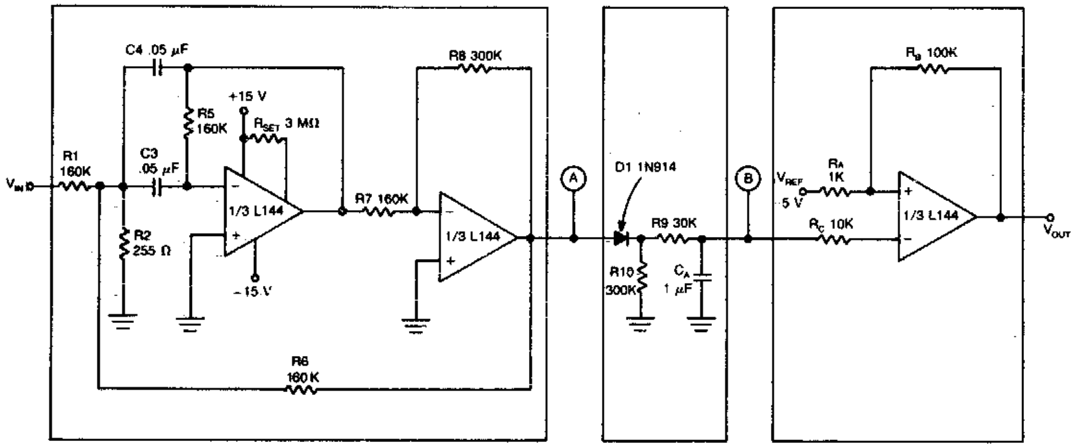
All resistor values are in ohms.

Fig. 25-15

Circuit Notes

This circuit provides for accurate half wave rectification of the incoming signal. For positive signals, the gain is 0; for negative signals, the gain is −1. By reversing both diodes, the polarity can be inverted. This circuit provides an accurate output, but the output impedance differs for the two input polarities and buffering may be needed. The output must slew through two diode drops when the input polarity reverses. The NE5535 device will work up to 10 kHz with less ttan 5% distortion.

TONE DETECTOR


C4 .05 μF
R5 160K
+15 V
R_{SET} 3 MΩ
R8 300K
R_B 100K
R1 160K
C3 .05 μF
V_{IN}
R2 255 Ω
1/3 L144
-15 V
R7 160K
1/3 L144
A
D1 1N914
R9 30K
R10 300K
C_A 1 μF
B
V_{REF} 5 V
R_A 1K
R_C 10K
1/3 L144
V_{OUT}
R6 160 K

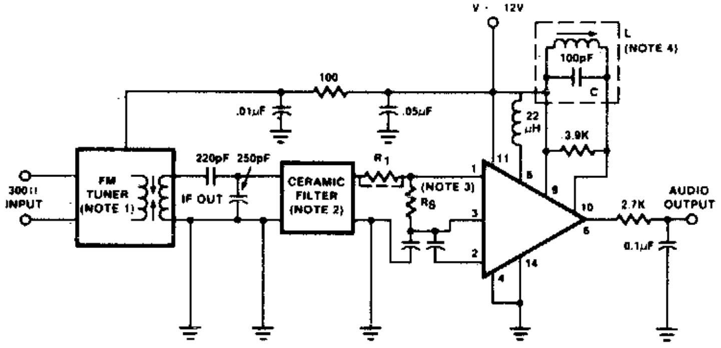
P_D = 7.5 mW

Fig. 25-16

Circuit Notes

The detector circuit is made up a two-amplifier multiple feedback bandpass filter followed by an ac-to-dc detector section and a Schmitt Trigger. The bandpass filter (with a Q of greater than 100) passes only 500 Hz inputs whch are in turn rectified by D1 and filtered by R9 and C_A. This filtering action in combination with the trigger level of 5 V for the Schmitt device insures that at least 55 cycles of 500 Hz input must be present before the output will react to a tone input.

FM TUNER WITH A SINGLE-TUNED DETECTOR COIL


V · 12V
L (NOTE 4)
100pF
C
100
.01μF
.05μF
22 μH
3.9K
300Ω INPUT
FM TUNER (NOTE 1)
220pF
250pF
IF OUT
CERAMIC FILTER (NOTE 2)
R_1
(NOTE 3)
R_S
1
2
3
4
5
6
9
10
11
14
2.7K
AUDIO OUTPUT
0.1μF

NOTES

All resistors values are typical and in ohms.

1. Weller 4SN3FIC or equivalent
2. Murate SFG 10.7mA or equivalent
3. R_S will affect stability depending on circuit layout. To increase stability R_S is decreased. Range of R_S is 330 to 50Ω, $R_1 + R_S \leq 330Ω$
4. L tunes with 100pF (C) at 10.7MHz Q_O unloaded ≈ 75 (G.I. EX27825 or equivalent)

Performance data at f_o = 98MHz, f_{MOD} = 400Hz, deviation =

± 74kHz:

−3dB limiting sensitivity	2μV (antenna level)
20dB quieting sensitivity	1μV (antenna level)
30dB quieting sensitivity	1.5μV (antenna level)

Fig. 25-17

MISSING PULSE DETECTOR

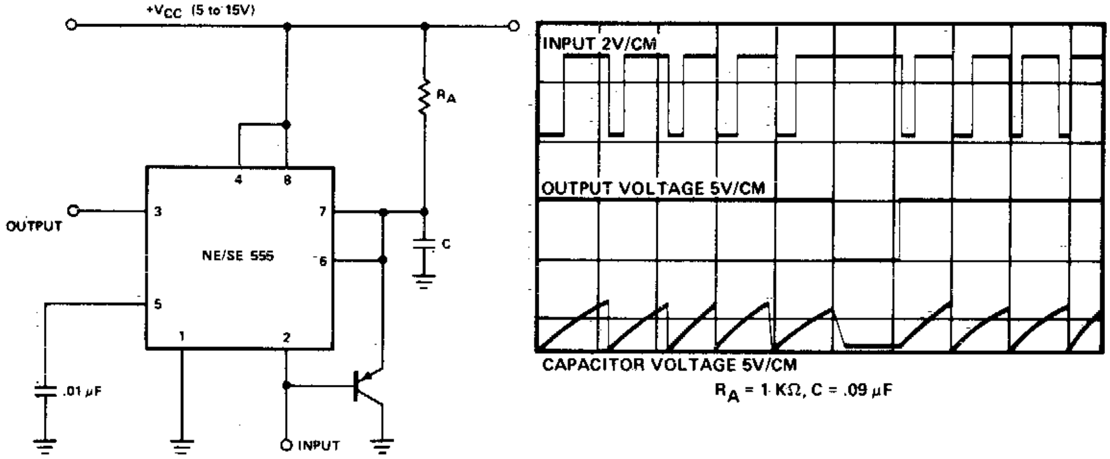

+VCC (5 to 15V)
RA
4
8
7
6
3
5
1
2
NE/SE 555
OUTPUT
C
.01 µF
INPUT
INPUT 2V/CM
OUTPUT VOLTAGE 5V/CM
CAPACITOR VOLTAGE 5V/CM
RA = 1 KΩ, C = .09 µF

Fig. 25-18

Circuit Notes

The timing cycle is continuously reset by the input pulse train. A change in frequency, or a missing pulse, allows completion of the timing cycle which causes a change in the output level. For this application, the time delay should be set to be slightly longer than the normal time between pulses. The graph shows the actual waveforms seen in this mode of operation.

HIGH SPEED PEAK DETECTOR

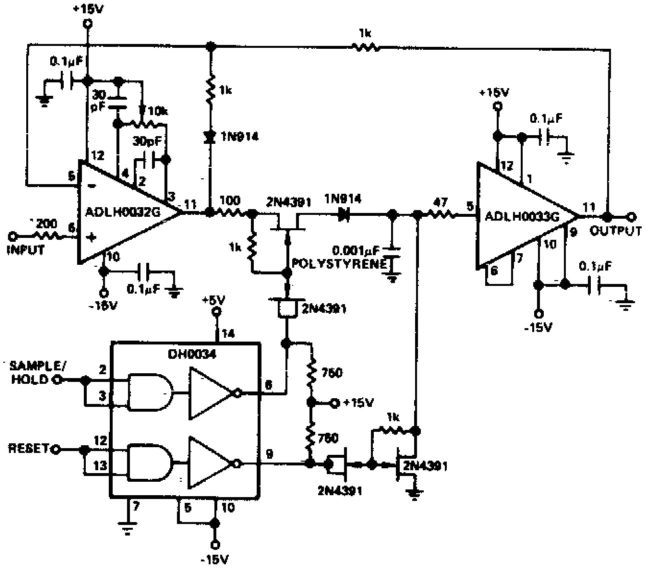
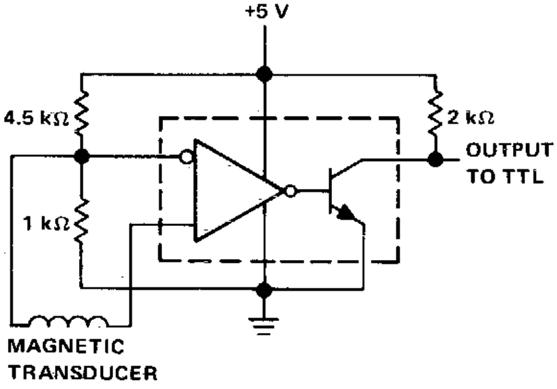

+15V
1k
0.1µF
30 pF
10k
30pF
1k
1N914
ADLH0032G
200
INPUT
-15V
0.1µF
100
2N4391
1N914
1k
0.001µF POLYSTYRENE
47
ADLH0033G
+15V
0.1µF
OUTPUT
-15V
0.1µF
2N4391
+5V
DH0034
SAMPLE/ HOLD
RESET
750
+15V
750
1k
2N4391
2N4391
-15V

Fig. 25-19

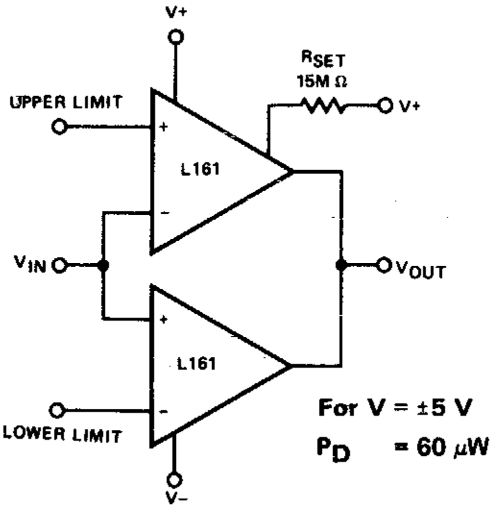
DETECTOR FOR MAGNETIC TRANSDUCER



+5 V
4.5 kΩ
2 kΩ
OUTPUT
TO TTL
1 kΩ
MAGNETIC
TRANSDUCER

Fig. 25-20

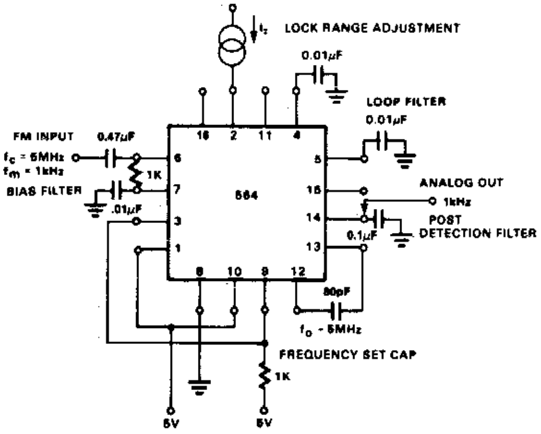
DOUBLE-ENDED LIMIT DETECTOR



V+
R_{SET}
15M Ω
V+
UPPER LIMIT
+
–
L161
V_{IN}
V_{OUT}
+
L161
–
LOWER LIMIT
V–
For V = ±5 V
P_D = 60 μW

Fig. 25-21

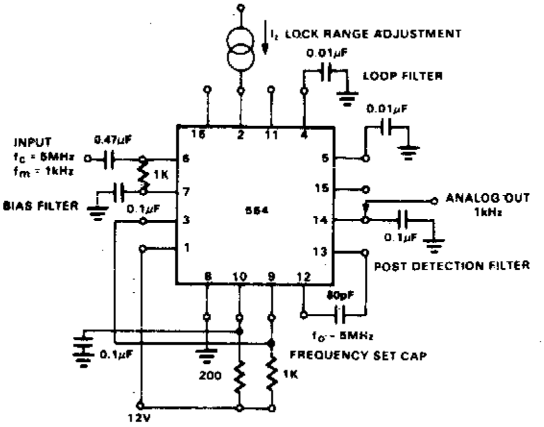
FM DEMODULATOR AT 5 V



I_2
LOCK RANGE ADJUSTMENT
0.01μF
LOOP FILTER
0.01μF
FM INPUT
0.47μF
f_c = 5MHz
f_m = 1kHz
1K
BIAS FILTER
.01μF
16
2
11
4
6
5
7
564
15
3
14
1
13
8
10
9
12
ANALOG OUT
1kHz
POST
DETECTION FILTER
0.1μF
80pF
f_o - 5MHz
FREQUENCY SET CAP
1K
5V
5V

Fig. 25-22

FM DEMODULATOR AT 12 V



I_2 LOCK RANGE ADJUSTMENT
0.01μF
LOOP FILTER
0.01μF
INPUT
0.47μF
f_c = 5MHz
f_m = 1kHz
1K
BIAS FILTER
0.1μF
16
2
11
4
6
5
7
15
564
3
14
1
13
8
10
9
12
ANALOG OUT
1kHz
0.1μF
POST DETECTION FILTER
80pF
f_o - 5MHz
0.1μF
FREQUENCY SET CAP
200
1K
12V

Fig. 25-23

PRECISION FULL WAVE RECTIFIER

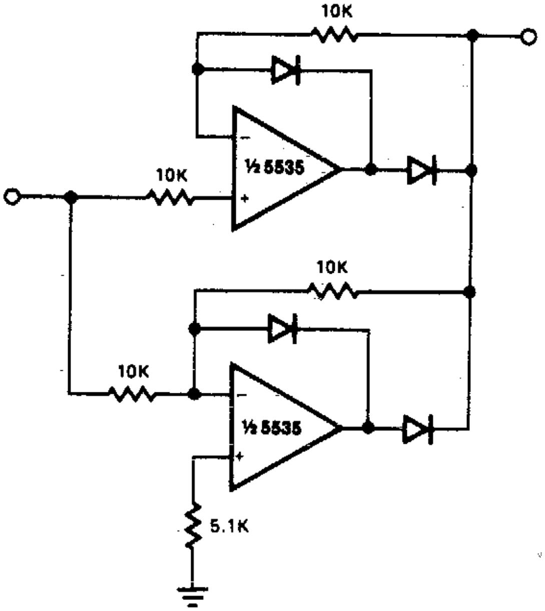

10K
½5535
10K
10K
10K
½5535
5.1K

Fig. 25-24

Circuit Notes

The circuit provides accurate full wave rectification. The output impedance is low for both input polarities, and the errors are small at all signal levels. Note that the output will not sink heavy current, except a small amount through the 10 K resistors. Therefore, the load applied should be referenced to ground or a negative voltage. Reversal of all diode polarities will reverse the polarity of the output. Since the outputs of the amplifiers must slew through two diode drops when the input polarity changes, 741 type devices give 5% distortion at about 300 Hz.

NEGATIVE PEAK DETECTOR

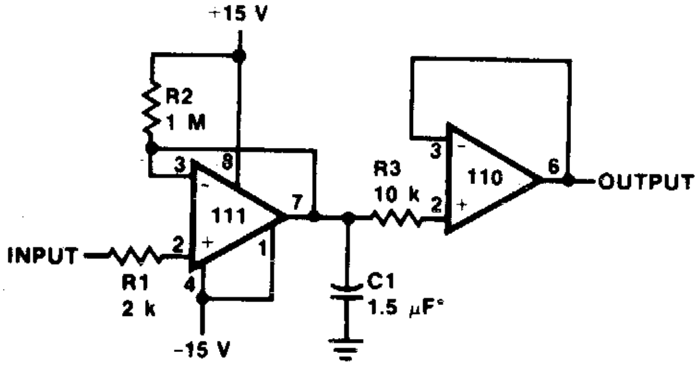
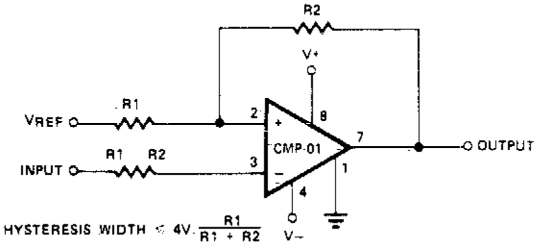

+15 V
R2
1 M
3
8
111
7
INPUT
2
1
R1
2 k
4
-15 V
R3
10 k
3
2
110
6
OUTPUT
C1
1.5 μF*

Fig. 25-25

*Solid tantalum

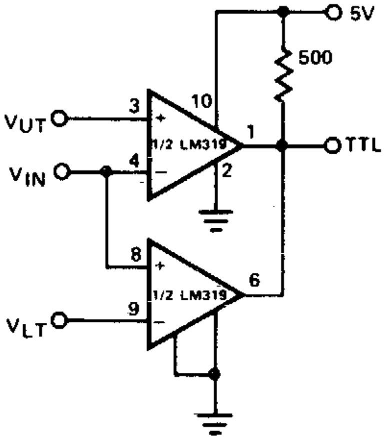
LEVEL DETECTOR WITH HYSTERESIS (POSITIVE FEEDBACK)


R2
V+
R1
VREF
2
+
8
CMP-01
7
OUTPUT
R1 R2
INPUT
3
−
1
4
V−

HYSTERESIS WIDTH $\leq 4V \cdot \frac{R1}{R1 + R2}$

Fig. 25-26

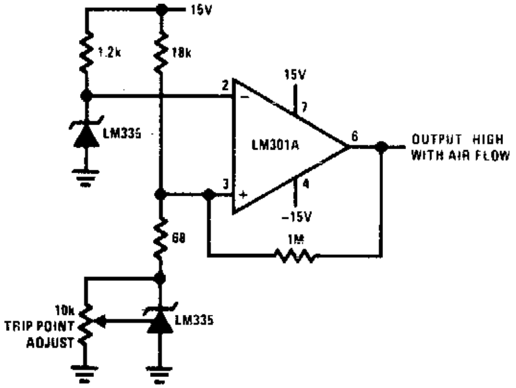
WINDOW DETECTOR


5V
500
VUT
3
+
10
1/2 LM319
1
TTL
VIN
4
−
2
8
+
1/2 LM319
6
VLT
9
−

$V_{OUT} = 5V$ for $V_{LT} < V_{IN} < V_{UT}$
$V_{OUT} = 0$ for $V_{IN} < V_{LT}$ or $V_{IN} > V_{UT}$

Fig. 25-27

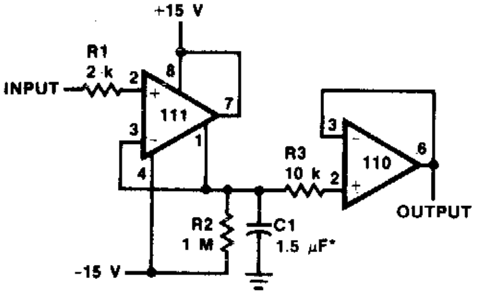
AIR FLOW DETECTOR


15V
1.2k
18k
15V
2
−
7
LM335
LM301A
6
OUTPUT HIGH WITH AIR FLOW
3
+
4
−15V
68
1M
10k
TRIP POINT ADJUST
LM335

*Self heating is used to detect air flow

Fig. 25-28

POSITIVE PEAK DETECTOR


+15 V
R1
2 k
INPUT
2
+
8
111
7
3
−
1
4
3
−
110
6
R3
10 k
2
+
OUTPUT
R2
1 M
C1
1.5 µF*
−15 V

*Solid tantalum

Fig. 25-29

26

Digital-to-Analog Converters

The sources of the following circuits are contained in the Sources section beginning on page 730. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

14-Bit Binary D/A Converter (Unipolar)
10-Bit D/A Converter
Fast Voltage Output D/A Converter
Resistor Terminated DAC (0 to −5 V Output)
Three-Digit BCD D/A Converter
8-Bit D/A Converter
High-Speed 8-Bit D/A Converter
10-Bit, 4 Quadrant Multiplexing D/A Converter (Offset Binary Coding)
8-Bit D/A Converter
±10 V Full-Scale Bipolar DAC
Precision 12-Bit D/A Converter
8-Bit D/A with Output Current-to-Voltage Conversion
16-Bit Binary DAC
±10 V Full-Scale Unipolar DAC
High-Speed Voltage Output DAC

14-BIT BINARY D/A CONVERTER (UNIPOLAR)

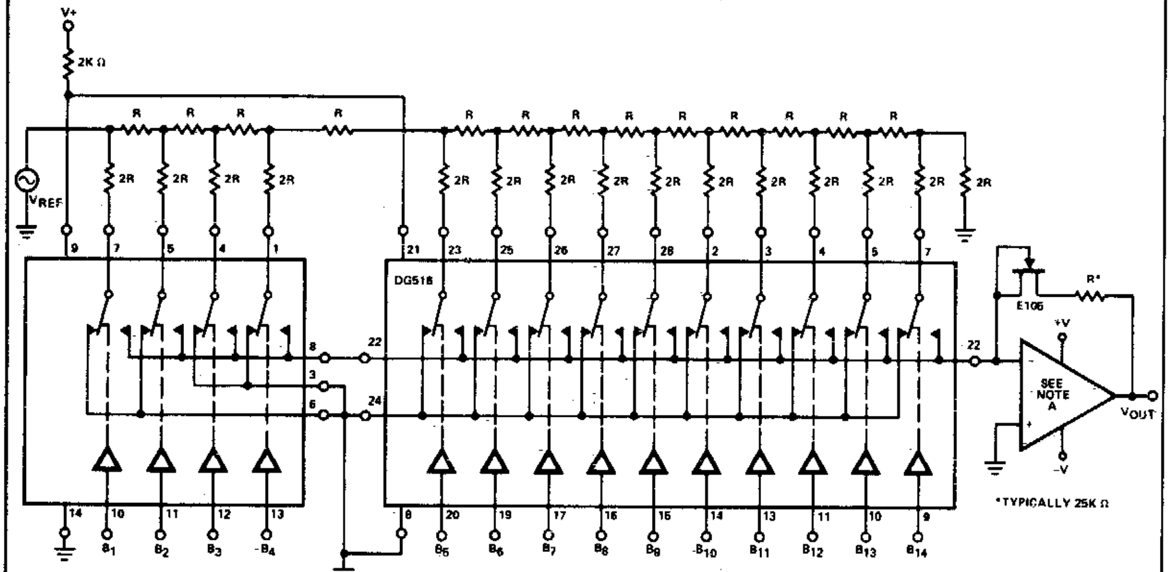

V+
2K Ω
R
2R
VREF
9
7
5
4
1
21
23
25
26
27
28
2
3
4
5
7
DG518
8
22
3
6
24
22
E106
R*
+V
-V
SEE NOTE A
VOUT
*TYPICALLY 25K Ω
14
10
11
12
13
8
20
19
17
16
15
14
13
11
10
9
B1
B2
B3
B4
B5
B6
B7
B8
B9
B10
B11
B12
B13
B14
BINARY INPUT

Figure 7

NOTE:

A. Op-Amp characteristics effect D/A accuracy and settling time. The following Op-Amps, listed in order of increasing speed, are suggested:

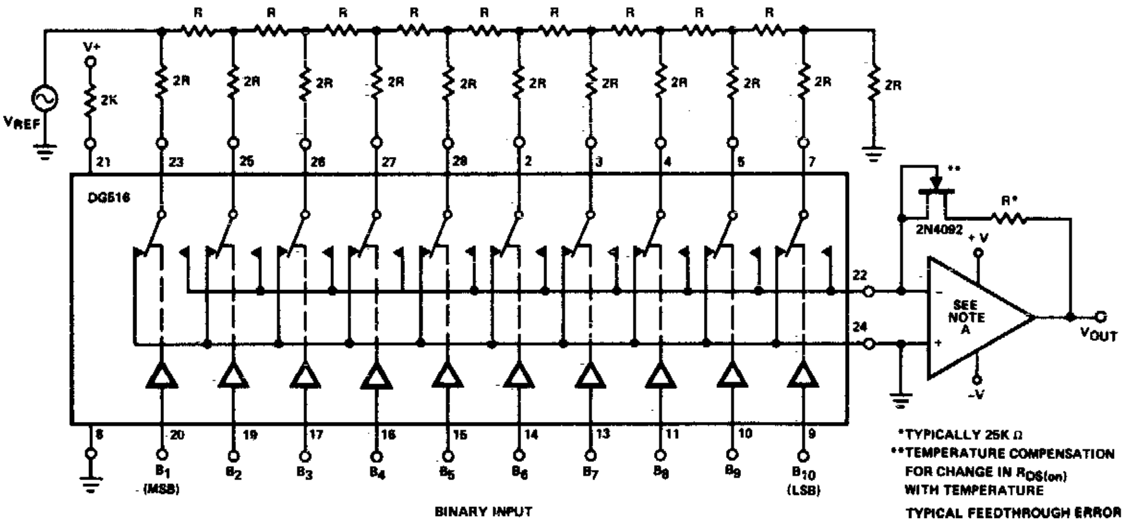
1. LM101A
2. LF156A
3. LM118

Unipolar Binary Operation

DIGITAL INPUT	ANALOG OUTPUT
1 1 1 1 1 1 1 1 1 1 1 1 1 1	$-V_{REF}(1-2^{-14})$
1 0 0 0 0 0 0 0 0 0 0 0 0 1	$-V_{REF}(1/2+2^{-14})$
1 0 0 0 0 0 0 0 0 0 0 0 0 0	$-V_{REF}/2$
0 1 1 1 1 1 1 1 1 1 1 1 1 1	$-V_{REF}(1/2-2^{-14})$
0 0 0 0 0 0 0 0 0 0 0 0 0 1	$-V_{REF}(2^{-14})$
0 0 0 0 0 0 0 0 0 0 0 0 0 0	0

Fig. 26-1

10 BIT D/A CONVERTER


R
2R
V+
2K
VREF
21
23
25
26
27
28
2
3
4
5
7
DG516
22
24
2N4092
R*
+V
SEE NOTE A
-V
VOUT
8
20
19
17
16
15
14
13
11
10
9
B1 (MSB)
B2
B3
B4
B5
B6
B7
B8
B9
B10 (LSB)
BINARY INPUT

*TYPICALLY 25K Ω
**TEMPERATURE COMPENSATION FOR CHANGE IN $R_{DS(on)}$ WITH TEMPERATURE

TYPICAL FEEDTHROUGH ERROR OF 10 mV p-p FOR V_{REF} = 10 V p-p, f = 100 kHz AND R = 25K Ω

Unipolar Binary Operation

DIGITAL INPUT	ANALOG OUTPUT
1 1 1 1 1 1 1 1 1 1	$-V_{REF}\ (1-2^{-10})$
1 0 0 0 0 0 0 0 0 1	$-V_{REF}\ (1/2+2^{-10})$
1 0 0 0 0 0 0 0 0 0	$-V_{REF}/2$
0 1 1 1 1 1 1 1 1 1	$-V_{REF}\ (1/2-2^{-10})$
0 0 0 0 0 0 0 0 0 1	$-V_{REF}\ (2^{-10})$
0 0 0 0 0 0 0 0 0 0	0

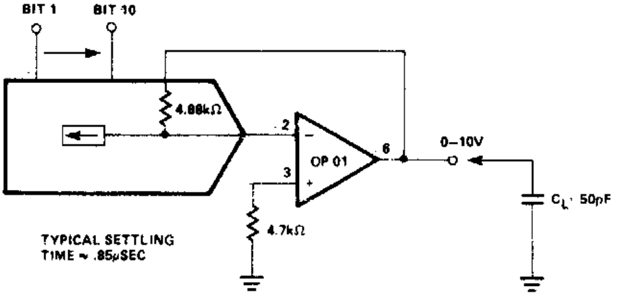
NOTE:

Op-Amp characteristics effect D/A accuracy and settling time. The following Op-Amps, listed in order of increasing speed, are suggested:

1. LM101A
2. LF156A
3. LM118

Fig. 26-2

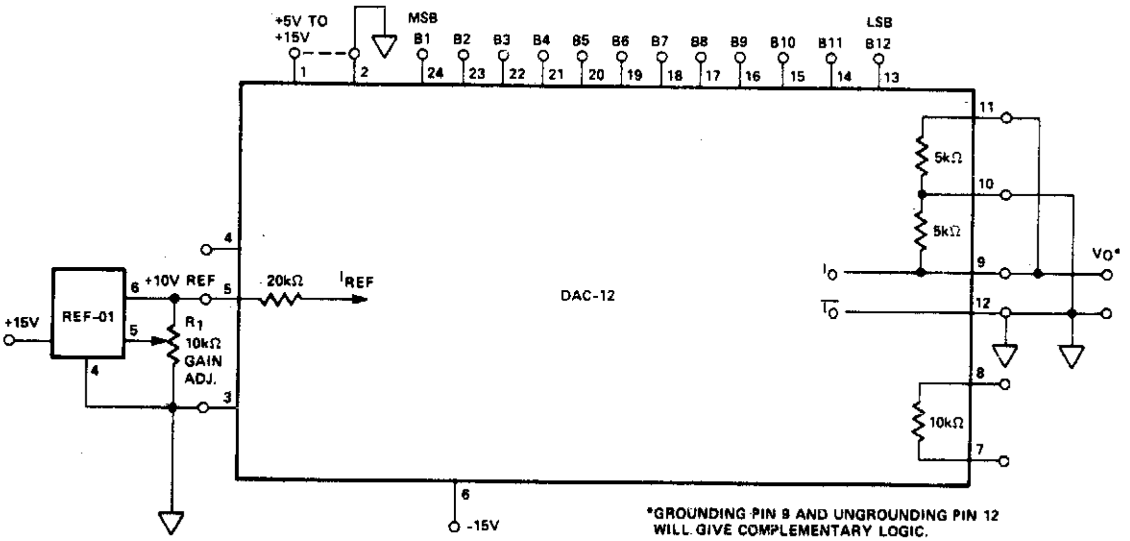
FAST VOLTAGE OUTPUT D/A CONVERTER


BIT 1
BIT 10
4.88kΩ
2
3
OP 01
6
0–10V
C_L 50pF
4.7kΩ

TYPICAL SETTLING TIME ~ .85μSEC

Fig. 26-3

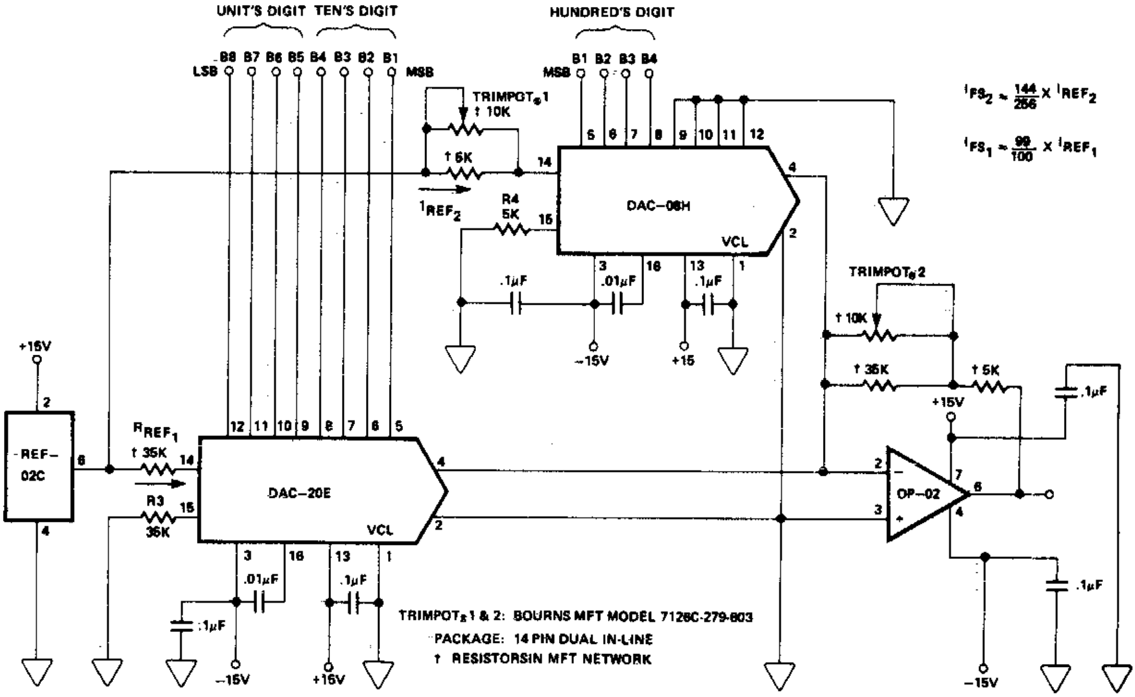
RESISTOR TERMINATED DAC (0 TO −5 V OUTPUT)


+5V TO +15V
MSB
LSB
B1 B2 B3 B4 B5 B6 B7 B8 B9 B10 B11 B12
1 2 24 23 22 21 20 19 18 17 16 15 14 13
REF-01
+15V
+10V REF
R1 10kΩ GAIN ADJ.
20kΩ
IREF
DAC-12
5kΩ
5kΩ
10kΩ
IO
IO
VO*
11 10 9 12 8 7
4 5 3 6
−15V

*GROUNDING PIN 9 AND UNGROUNDING PIN 12 WILL GIVE COMPLEMENTARY LOGIC.

Fig. 26-4

THREE-DIGIT BCD D/A CONVERTER


UNIT'S DIGIT
TEN'S DIGIT
HUNDRED'S DIGIT
LSB B8 B7 B6 B5 B4 B3 B2 B1 MSB
MSB B1 B2 B3 B4
TRIMPOT® 1 † 10K
† 6K
IREF2
R4 5K
DAC-08H
VCL
.1μF
.01μF
.1μF
−15V
+15
IFS2 ≈ 144/256 × IREF2
IFS1 ≈ 99/100 × IREF1
TRIMPOT® 2
† 10K
† 35K
† 5K
.1μF
+15V
OP-02
+15V
REF-02C
RREF1 † 35K
R3 35K
DAC-20E
VCL
.01μF
.1μF
.1μF
−15V
+15V
−15V
.1μF

TRIMPOTS 1 & 2: BOURNS MFT MODEL 7126C-279-803

PACKAGE: 14 PIN DUAL IN-LINE

† RESISTORSIN MFT NETWORK

Fig. 26-5

8-BIT D/A CONVERTER

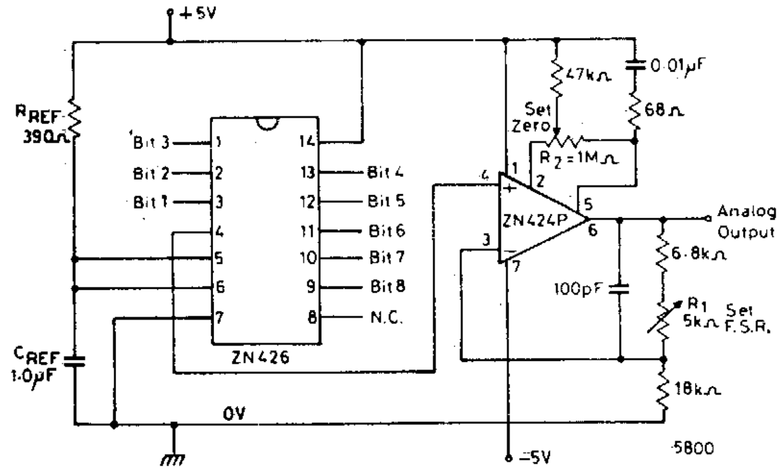
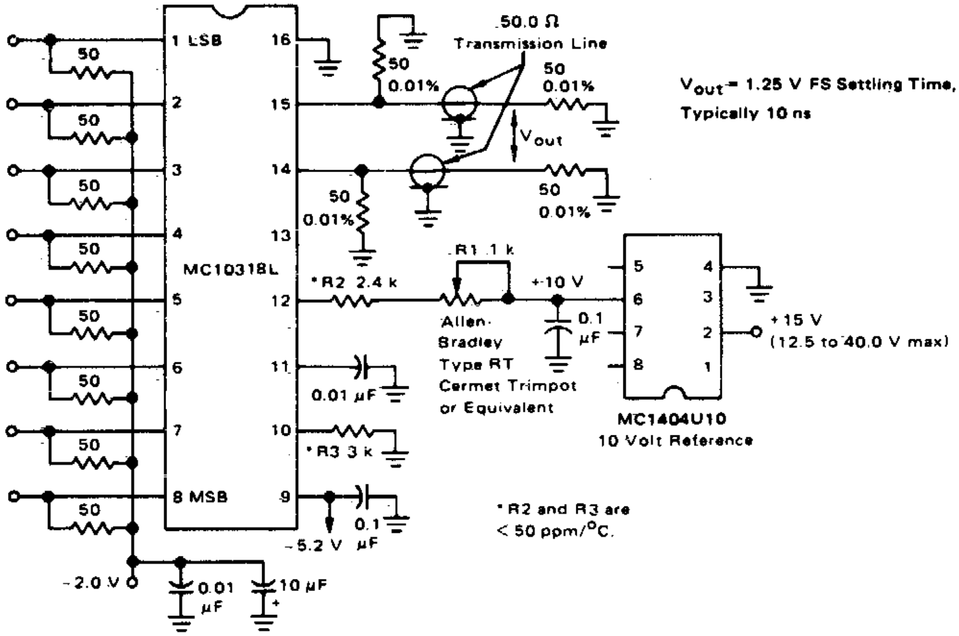

+5V
R_REF 390Ω
C_REF 1.0μF
Bit 3
Bit 2
Bit 1
1
2
3
4
5
6
7
14
13
12
11
10
9
8
Bit 4
Bit 5
Bit 6
Bit 7
Bit 8
N.C.
ZN426
0V
47kΩ
0.01μF
68Ω
Set Zero
R2 = 1MΩ
ZN424P
Analog Output
6.8kΩ
R1 5kΩ
Set F.S.R.
100pF
18kΩ
−5V
5800

Fig. 26-6

HIGH-SPEED 8-BIT D/A CONVERTER

I_{FS} is set to 51.000 mA with R1


1 LSB
2
3
4
5
6
7
8 MSB
16
15
14
13
12
11
10
9
50
MC10318L
50.0 Ω Transmission Line
50 0.01%
V_out
*R2 2.4 k
R1 1 k
Allen-Bradley Type RT Cermet Trimpot or Equivalent
+10 V
0.1 μF
0.01 μF
*R3 3 k
0.1 μF
−5.2 V
−2.0 V
0.01 μF
10 μF
MC1404U10
10 Volt Reference
+15 V
(12.5 to 40.0 V max)

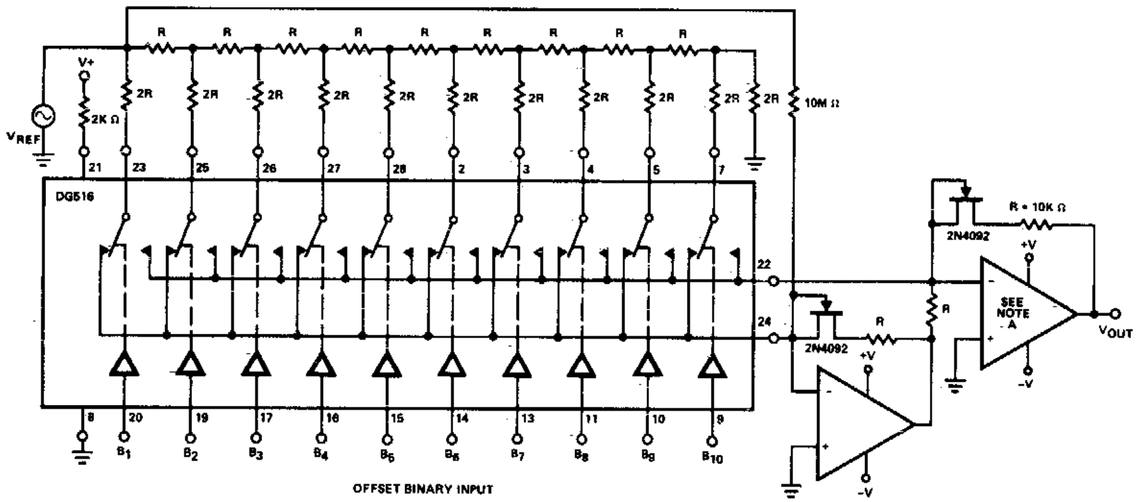
Inputs are MECL 10K Compatible

V_{out} = 1.25 V FS Settling Time, Typically 10 ns

*R2 and R3 are < 50 ppm/°C.

Fig. 26-7

10-BIT, 4 QUADRANT MULTIPLEXING D/A CONVERTER (OFFSET BINARY CODING)



R
2R
10M Ω
V+
2K Ω
VREF
DG516
21 23 25 26 27 28 2 3 4 5 7
22
24
8 20 19 17 16 15 14 13 11 10 9
B1 B2 B3 B4 B5 B6 B7 B8 B9 B10
OFFSET BINARY INPUT
2N4092
R = 10K Ω
+V
-V
SEE NOTE A
VOUT

Bipolar (Offset Binary)* Operation

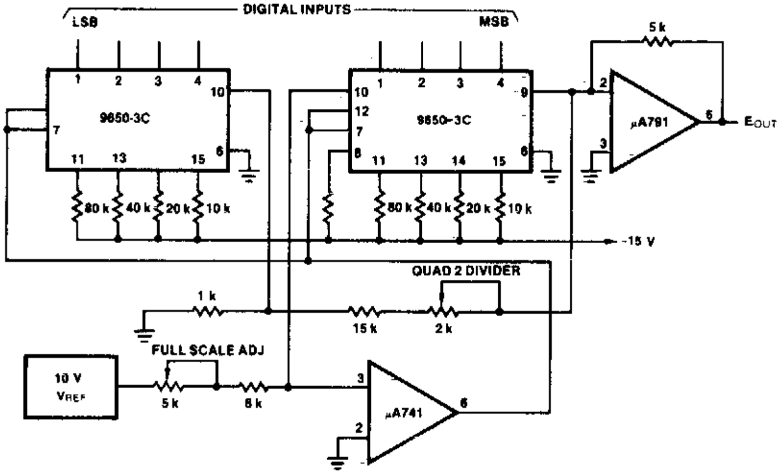
DIGITAL INPUT	ANALOG OUTPUT
1 1 1 1 1 1 1 1 1 1	$-V_{REF}\ (1 - 2^{-9})$
1 0 0 0 0 0 0 0 0 1	$-V_{REF}\ (2^{-9})$
1 0 0 0 0 0 0 0 0 0	0
0 1 1 1 1 1 1 1 1 1	$V_{REF}\ (2^{-9})$
0 0 0 0 0 0 0 0 0 1	$V_{REF}\ (1 - 2^{-9})$
0 0 0 0 0 0 0 0 0 0	V_{REF}

NOTE: 1 LSB = $2^{-9}\ V_{REF}$

*Complementing B_1 (MSB) will give 2's complement coding.

Fig. 26-8

8-BIT D/A CONVERTER

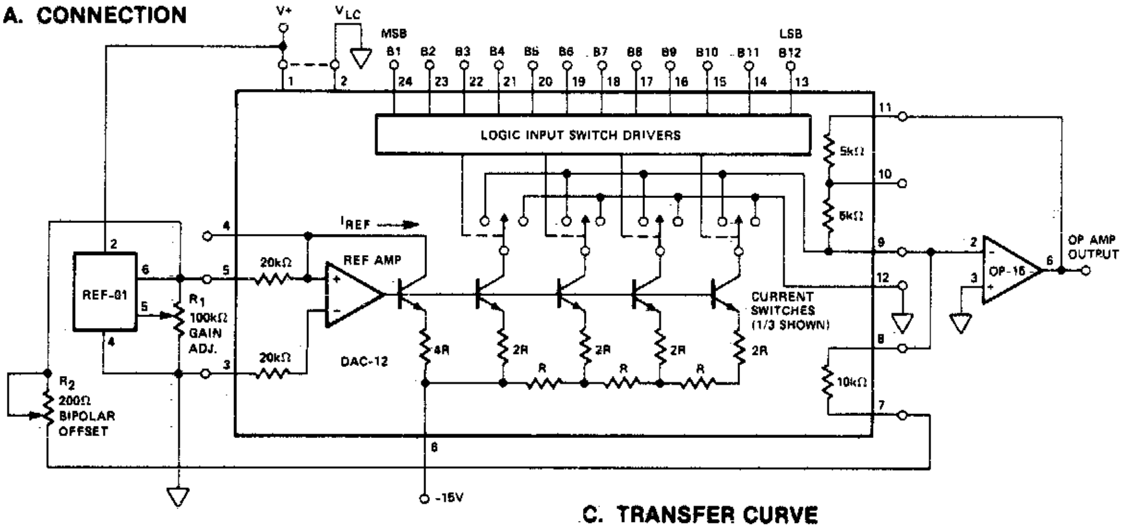


DIGITAL INPUTS
LSB
MSB
9650-3C
9650-3C
1 2 3 4 10 7 11 13 15 6
10 1 2 3 4 9 12 7 8 11 13 14 15 6
80 k 40 k 20 k 10 k
80 k 40 k 20 k 10 k
-15 V
5 k
µA791
2 3 6
EOUT
QUAD 2 DIVIDER
1 k
15 k
2 k
FULL SCALE ADJ
10 V VREF
5 k
8 k
µA741
3 2 6

Fig. 26-9

±10 V FULL-SCALE BIPOLAR DAC

A. CONNECTION


V+
V_LC
MSB
LSB
B1 B2 B3 B4 B5 B6 B7 B8 B9 B10 B11 B12
LOGIC INPUT SWITCH DRIVERS
5kΩ
5kΩ
10kΩ
I_REF
REF AMP
20kΩ
20kΩ
REF-01
R1 100kΩ GAIN ADJ.
R2 200Ω BIPOLAR OFFSET
DAC-12
4R
2R
R
CURRENT SWITCHES (1/3 SHOWN)
OP-16
OP AMP OUTPUT
-15V

B. ADJUSTMENT PROCEDURE

1. SET ALL BITS TO "0" STATE
2. ADJUST R_2 UNTIL OUTPUT IS −10V
3. BIT 1 to "1" STATE
4. ADJUST R_1 UNTIL OUTPUT = 0.00V

C. TRANSFER CURVE

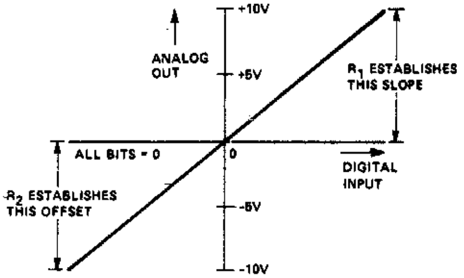


Fig. 26-10

PRECISION 12-BIT D/A CONVERTER

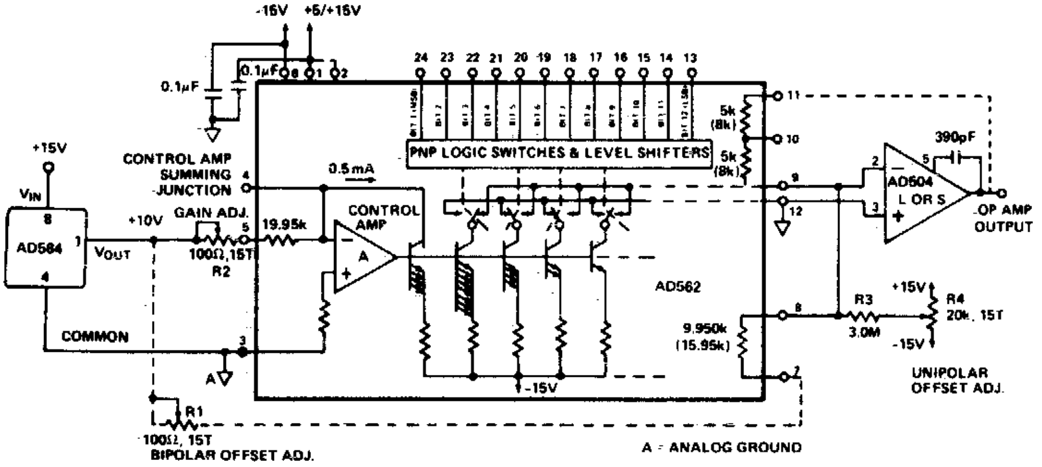

-15V
+5/+15V
0.1μF
0.1μF
24 23 22 21 20 19 18 17 16 15 14 13
PNP LOGIC SWITCHES & LEVEL SHIFTERS
5k (8k)
5k (8k)
390pF
AD504 L OR S
OP AMP OUTPUT
+15V
V_IN
AD584
V_OUT
COMMON
CONTROL AMP SUMMING JUNCTION
GAIN ADJ.
+10V
100Ω,15T R2
19.95k
0.5mA
CONTROL AMP
A
AD562
9.950k (15.95k)
R3 3.0M
R4 20k, 15T
+15V
-15V
UNIPOLAR OFFSET ADJ.
-15V
R1
100Ω, 15T BIPOLAR OFFSET ADJ.
A = ANALOG GROUND

Fig. 26-11

8-BIT D/A WITH OUTPUT CURRENT-TO-VOLTAGE CONVERSION

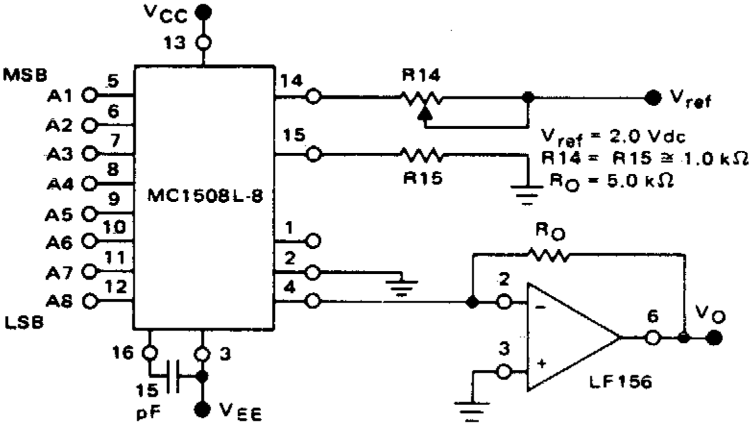

VCC
13
MSB
A1
A2
A3
A4
A5
A6
A7
A8
LSB
5
6
7
8
9
10
11
12
MC1508L-8
14
15
1
2
4
16
3
15 pF
VEE
R14
R15
Vref
Vref = 2.0 Vdc
R14 = R15 ≅ 1.0 kΩ
RO = 5.0 kΩ
RO
2
3
6
VO
LF156

Fig. 26-12

Theoretical V_O

$$V_O = \frac{V_{ref}}{R14}(R_O)\left[\frac{A1}{2} + \frac{A2}{4} + \frac{A3}{8} + \frac{A4}{16} + \frac{A5}{32} + \frac{A6}{64} + \frac{A7}{128} + \frac{A8}{256}\right]$$

Adjust V_{ref}, R14 or R_O so that V_O with all digital inputs at high level is equal to 9.961 volts.

$$V_O = \frac{2V}{1\,k}(5\,k)\left[\frac{1}{2} + \frac{1}{4} + \frac{1}{8} + \frac{1}{16} + \frac{1}{32} + \frac{1}{64} + \frac{1}{128} + \frac{1}{256}\right]$$

$$= 10\,V\left[\frac{255}{256}\right] = 9.961\,V$$

16-BIT BINARY DAC

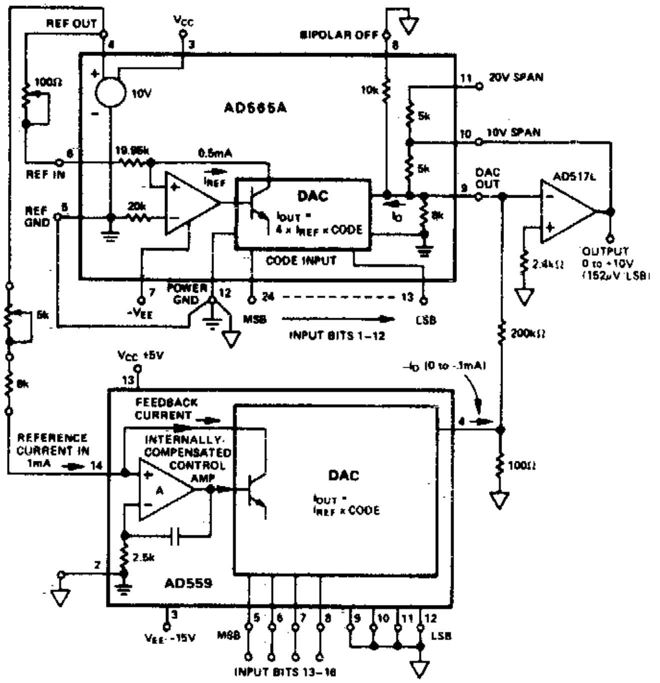
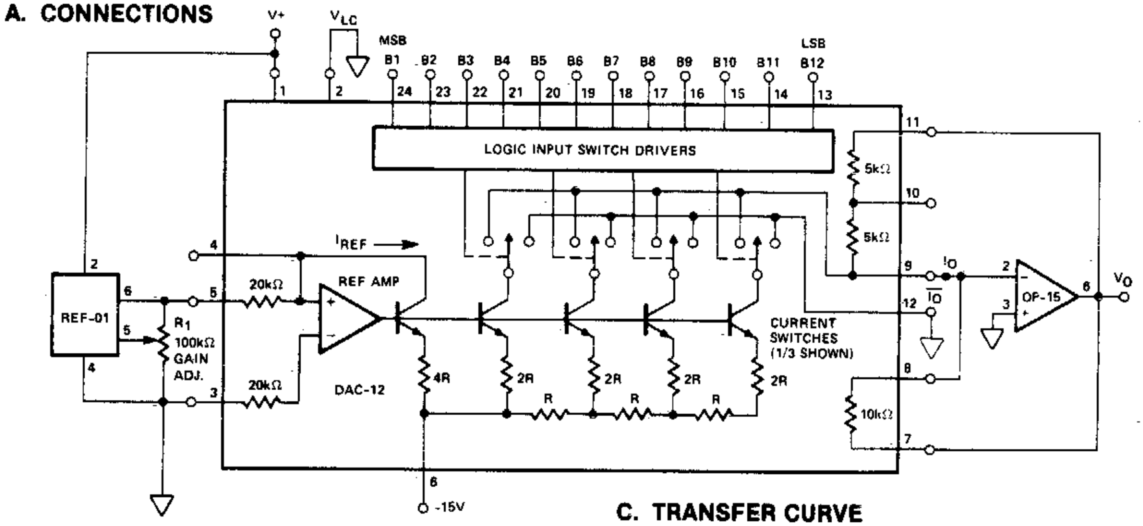

REF OUT
VCC
BIPOLAR OFF
4
3
8
100Ω
10V
AD565A
10k
11
20V SPAN
5k
10
10V SPAN
6
REF IN
19.95k
0.5mA
IREF
5k
DAC OUT
9
AD517L
REF GND
5
20k
DAC
IOUT = 4 × IREF × CODE
IO
8k
CODE INPUT
2.4kΩ
OUTPUT 0 to +10V (152μV LSB)
7
-VEE
POWER GND
12
24
13
MSB
LSB
INPUT BITS 1–12
5k
8k
200kΩ
VCC +5V
13
-IO (0 to -1mA)
FEEDBACK CURRENT
4
REFERENCE CURRENT IN 1mA
14
INTERNALLY-COMPENSATED CONTROL AMP
A
DAC
IOUT = IREF × CODE
100Ω
2.5k
2
AD559
3
VEE -15V
MSB
5
6
7
8
9
10
11
12
LSB
INPUT BITS 13–16

Fig. 26-13

±10 V FULL-SCALE UNIPOLAR DAC


A. CONNECTIONS

V+
V_LC
MSB
LSB
B1 B2 B3 B4 B5 B6 B7 B8 B9 B10 B11 B12
LOGIC INPUT SWITCH DRIVERS
REF-01
R1 100kΩ GAIN ADJ.
20kΩ
20kΩ
I_REF
REF AMP
DAC-12
4R
2R
R
CURRENT SWITCHES (1/3 SHOWN)
5kΩ
5kΩ
10kΩ
I_O
Ī_O
OP-15
V_O
-15V

B. ADJUSTMENT PROCEDURE

1. ALL BITS TO "1" STATE ("0" STATE IF PINS 9 AND 12 INTERCHANGED)
2. ADJUST R_1 UNTIL OUTPUT IS +9.9975

$$\frac{4095}{4096} \times 10V$$

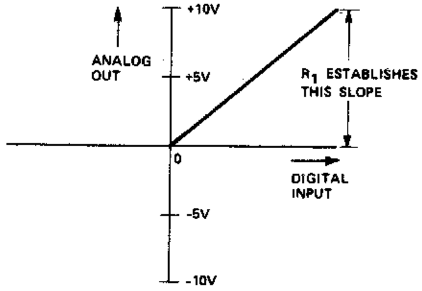
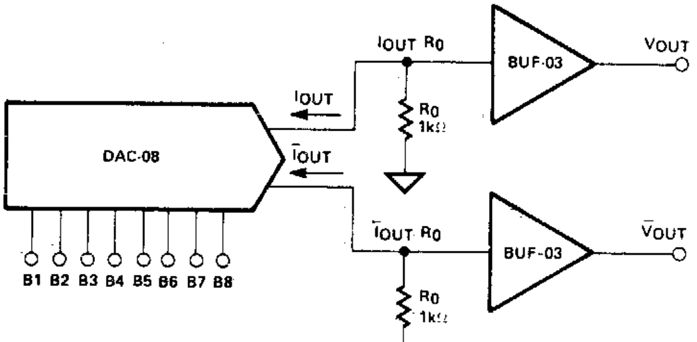

C. TRANSFER CURVE

Fig. 26-14

HIGH-SPEED VOLTAGE OUTPUT DAC



DAC-08
I_OUT
Ī_OUT
I_OUT R_O
Ī_OUT R_O
R_O 1kΩ
R_O 1kΩ
BUF-03
BUF-03
V_OUT
V̄_OUT
B1 B2 B3 B4 B5 B6 B7 B8

1/2 LSB SETTLING TIME 100nsec

SYSTEM WILL DRIVE CABLES OR TWISTED PAIRS.

Fig. 26-15

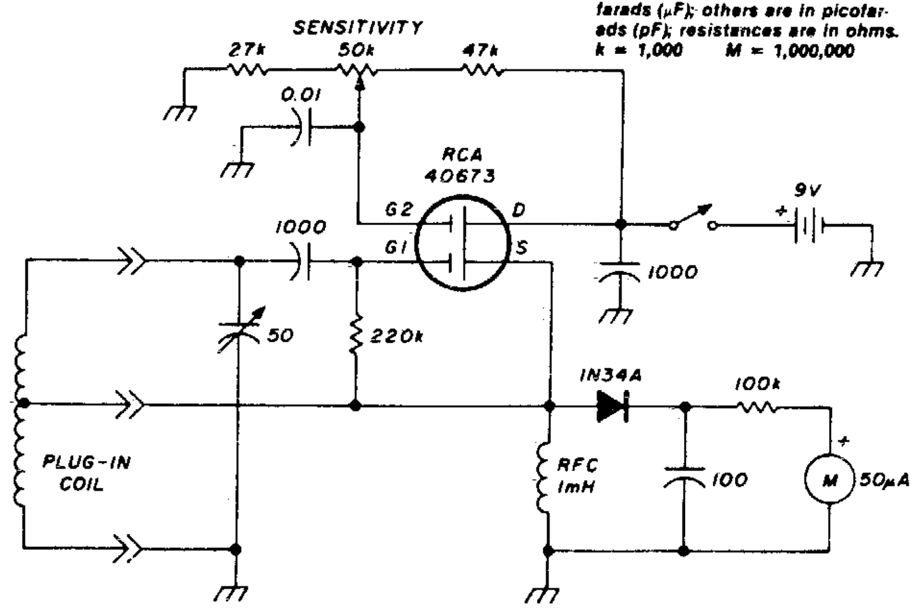
27

Dip Meters

The sources of the following circuits are contained in the Sources section beginning on page 730. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

- Dip Meter Using Dual-Gate IGFET (MOSFET)
- Varicap-Tuned FET DIP Meter with 1 kHz Modulator
- Dip Meter Using N-Channel IGFET (MOSFET) and Separate Diode Detector
- Basic Grid-Dip Meter
- Dip Meter Using Germanium PNP
- Bipolar Transistor with Separate Diode Detector
- Gate-Dip Meter Covers 1.8 - 150 MHz
- Dip Meter Using Silicon Junction FET

DIP METER USING DUAL-GATE IGFET (MOSFET)


SENSITIVITY
27k
50k
47k
0.01
RCA
40673
G2
D
G1
S
9V
+
1000
1000
50
220k
1N34A
100k
PLUG-IN
COIL
RFC
1mH
100
M
50μA
+

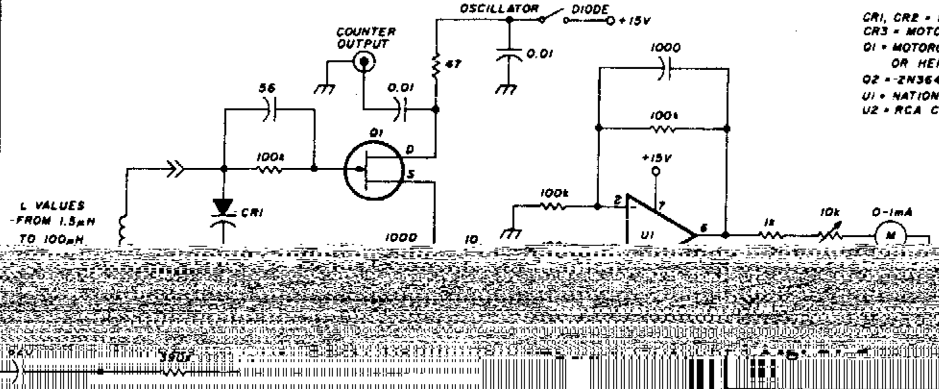
Except as indicated, decimal values of capacitance are in microfarads (μF); others are in picofarads (pF); resistances are in ohms.
k = 1,000 M = 1,000,000

Fig. 27-1

Circuit Notes

Gate 2 is used to adjust the oscillation level.

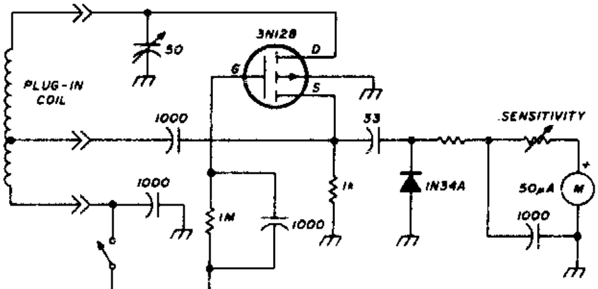
VARICAP-TUNED FET DIP METER WITH 1-kHz MODULATION


OSCILLATOR
DIODE
+15V
COUNTER
OUTPUT
47
0.01
56
0.01
1000
100k
100k
Q1
D
S
+15V
100k
2
7
6
U1
1k
10k
0-1mA
M
L VALUES
FROM 1.5μH
TO 100μH
CR1
1000
10

CR1, CR2 = MOTOROLA MVAM-2
CR3 = MOTOROLA MSD101
Q1 = MOTOROLA MPF102, HEP802,
OR HEP-F0015
Q2 = 2N3643
U1 = NATIONAL LM301A
U2 = RCA CD4001

Except as indicated, decimal values of capacitance are in microfarads (μF); others are in picofarads (pF); resistances are in ohms.
k = 1,000 M = 1,000,000

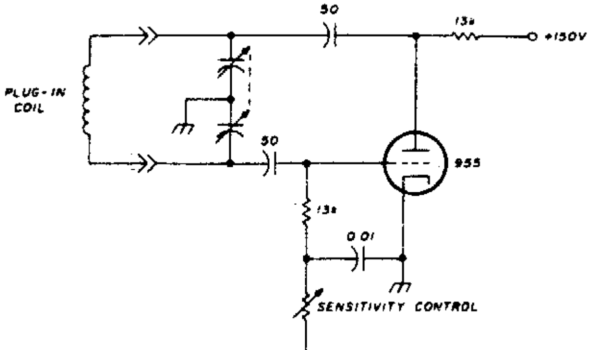
DIP METER USING N-CHANNEL IGFET (MOSFET) AND SEPARATE DIODE DETECTOR


3N128
50
D
G
S
PLUG-IN COIL
1000
33
SENSITIVITY
1000
1k
1N34A
50μA
M
+
1000
1M
1000
1000
+
9V

Except as indicated, decimal values of capacitance are in microfarads (μF); others are in picofarads (pF); resistances are in ohms. k = 1,000 M = 1,000,000

Fig. 27-3

BASIC GRID-DIP METER


50
13k
+150V
PLUG-IN COIL
50
955
13k
0.01
SENSITIVITY CONTROL
M
200μA

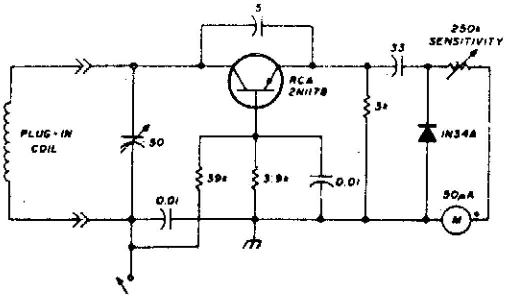
Except as indicated, decimal values of capacitance are in micro farads (μF); others are in picofarads (pF); resistances are in ohms. k = 1,000 M = 1,000,000

Circuit Notes

This circuit uses a triode vacuum-tube (9002 and 6C4 also commonly used).

Fig. 27-4

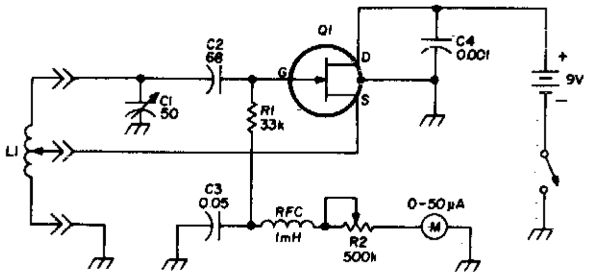
DIP METER USING GERMANIUM PNP BIPOLAR TRANSISTOR WITH SEPARATE DIODE DETECTOR


5
250k
SENSITIVITY
33
RCA 2N1178
3k
1N34A
PLUG-IN COIL
50
39k
3.9k
0.01
0.01
50μA
M
+
+
13.5V

Except as indicated, decimal values of capacitance are in microfarads (μF); others are in picofarads (pF); resistances are in ohms. k = 1,000 M = 1,000,000

Fig. 27-5

GATE-DIP METER COVERS 1.8 - 150 MHz


Q1
C2
68
G
D
S
C4
0.001
+
9V
−
C1
50
R1
33k
L1
C3
0.05
RFC
1mH
R2
500k
0-50µA
M

Coil data.

frequency range (MHz)	no. turns	wire size AWG	(mm)	winding length inches	(mm)	tap*	coil diameter inches	(mm)
1.8 - 3.8	82	26 enamel	(0.4)	1 9/16	(40.0)	12	1¼	(32)
3.6 - 7.3	29	26 enamel	(0.4)	9/16	(14.5)	5	1¼	(32)
7.3 - 14.4	18	22 enamel	(0.6)	3/4	(19.0)	3	1	(25)
14.4 - 32	7	22 enamel	(0.6)	1/2	(12.5)	2	1	(25)
29 - 64	3½	18 tinned	(1.0)	3/4	(19.0)	3/4	1	(25)

61 - 150 Hairpin of 16 no. AWG (1.3mm) wire, 5/8 inch (16mm) spacing, 2 3/8 inches (60mm) long including coil-form pins. Tapped at 2 inches (51mm) from ground end.

*Turns from ground-end. 1 inch (25mm) forms are Millen 45004 available from Burstein-Applebee

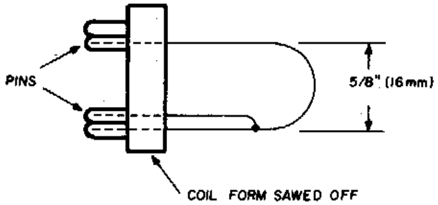

PINS
5/8" (16mm)
COIL FORM SAWED OFF

Fig. 27-6

DIP METER USING SILICON JUNCTION FET

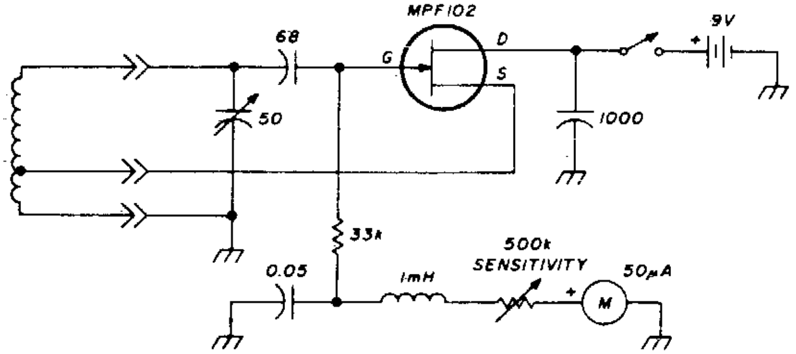

MPF102
68
G
D
S
9V
+
50
1000
33k
0.05
1mH
500k
SENSITIVITY
50µA
+
M

Fig. 27-7

Except as indicated, decimal values of capacitance are in microfarads (µF); others are in picofarads (pF); resistances are in ohms.
k = 1,000 M = 1,000,000

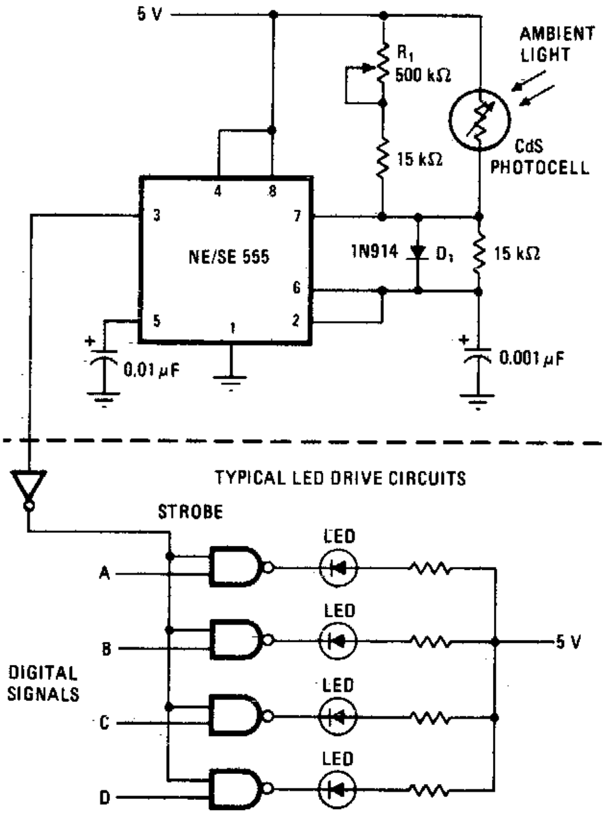
28

Displays

The sources of the following circuits are contained in the Sources section beginning on page 730. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

LED Brightness Control
LED Bar/Dot Level Meter
60 dB Dot Mode Display
Bar Display with Alarm Flasher
12-Hour Clock with Gas Discharge Displays
Precision Frequency Counter (~ 1 MHz Maximum)
Exclamation Point Display
LED Bar Peak Program Meter Display for Audio
10 MHz Universal Counter

LED BRIGHTNESS CONTROL



5 V
R₁
500 kΩ
AMBIENT LIGHT
CdS PHOTOCELL
15 kΩ
4
8
3
7
NE/SE 555
6
5
1
2
1N914
D₁
15 kΩ
0.01 μF
0.001 μF
TYPICAL LED DRIVE CIRCUITS
STROBE
LED
A
B
C
D
DIGITAL SIGNALS
5 V

Fig. 28-1

Circuit Notes

The brightness of LED display is varied by using a photocell in place of one timing resistor in a 555 timer, and bypassing the other timing resistor to boost the timer's maximum duty cycle. The result is a brighter display in sunlight and a fainter one in the dark.

LED BAR/DOT LEVEL METER

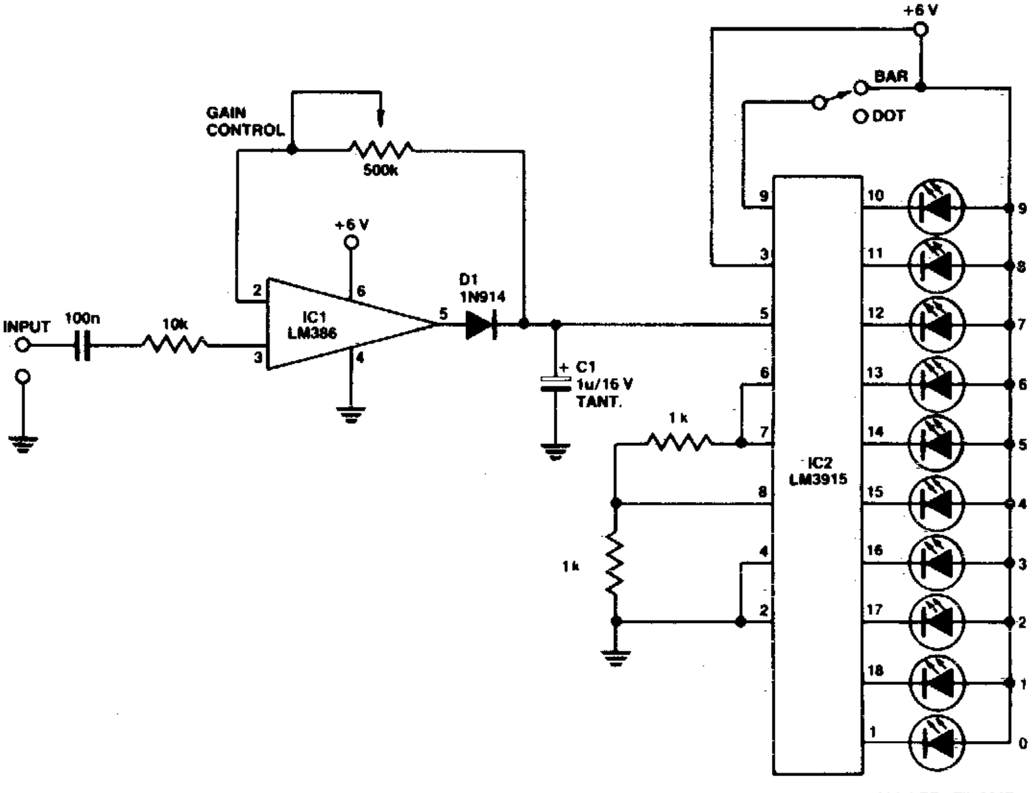
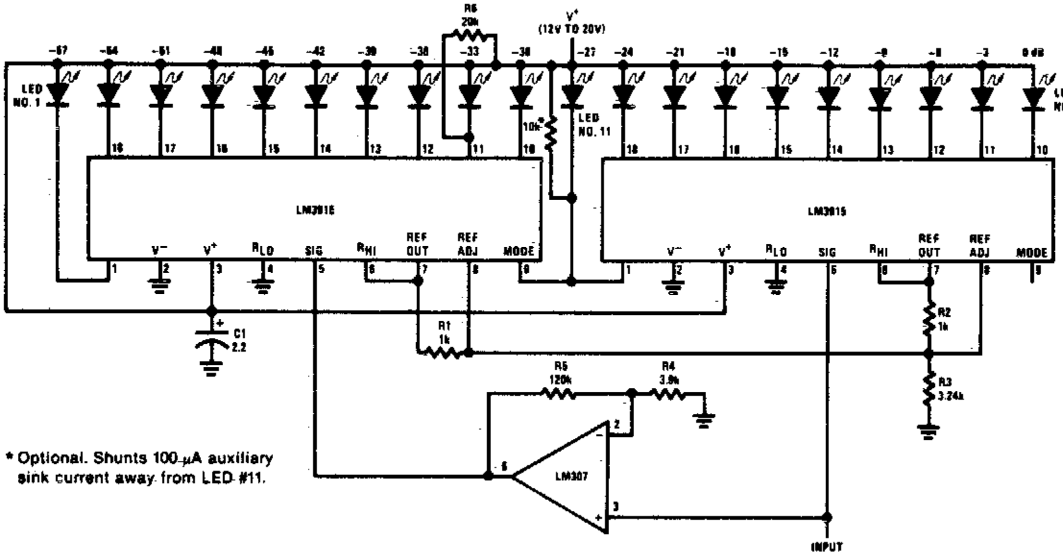

+6 V
BAR
DOT
GAIN CONTROL
500k
+6 V
D1
1N914
INPUT
100n
10k
IC1
LM386
C1
1u/16 V
TANT.
1 k
1 k
IC2
LM3915
ALL LEDs TIL 220R

Fig. 28-2

Circuit Notes

A simple level of power meter can be arranged to give a bar or dot display for a hi-fi system. Use green LEDs for 0 to 7; yellow for 8 and red for 9 to indicate peak power. The gain control is provided to enable calibration on the equipment with which the unit is used. Because the unit draws some 200 mA, a power supply is advisable instead of running the unit from batteries.

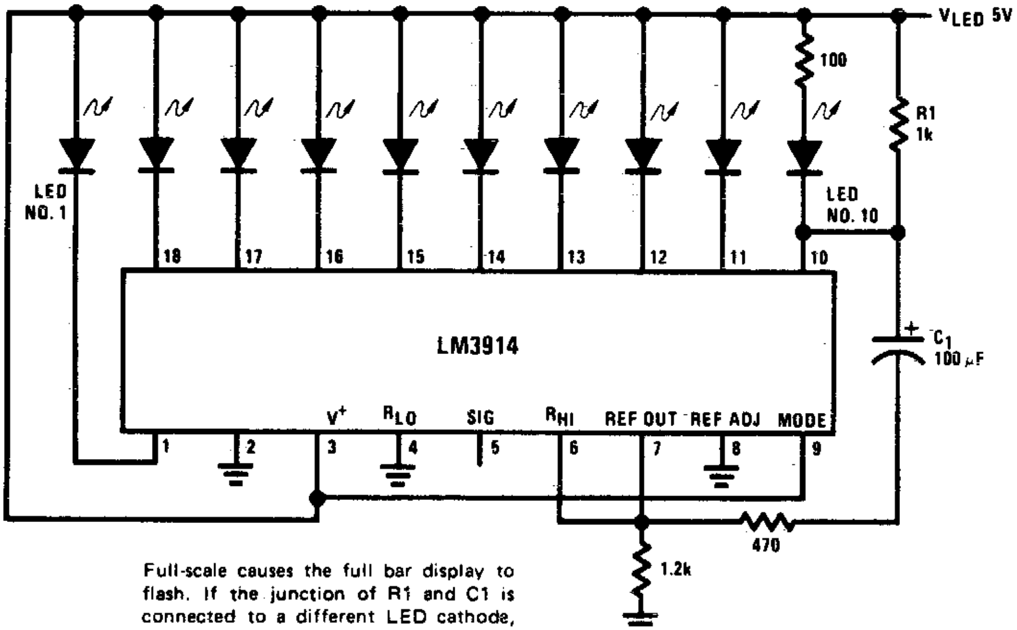
60 dB DOT MODE DISPLAY


R6
20k
V+
(12V TO 20V)
−57
−54
−51
−48
−45
−42
−39
−36
−33
−30
−27
−24
−21
−18
−15
−12
−9
−6
−3
0 dB
LED NO. 1
10k*
LED NO. 11
LED NO. 20
18
17
16
15
14
13
12
11
10
LM3915
V−
V+
RLO
SIG
RHI
REF OUT
REF ADJ
MODE
1
2
3
4
5
6
7
8
9
C1
2.2
R1
1k
R2
1k
R3
3.24k
R5
120k
R4
3.9k
LM307
6
2
3
INPUT
(10V FULL-SCALE)

* Optional. Shunts 100 μA auxiliary sink current away from LED #11.

Fig. 28-3

BAR DISPLAY WITH ALARM FLASHER


VLED 5V
100
R1
1k
LED NO. 1
LED NO. 10
18
17
16
15
14
13
12
11
10
LM3914
V+
RLO
SIG
RHI
REF OUT
REF ADJ
MODE
1
2
3
4
5
6
7
8
9
C1
100 μF
470
1.2k

Full-scale causes the full bar display to flash. If the junction of R1 and C1 is connected to a different LED cathode, the display will flash when that LED lights, and at any higher input signal.

Fig. 28-4

12-HOUR CLOCK WITH GAS DISCHARGE DISPLAYS

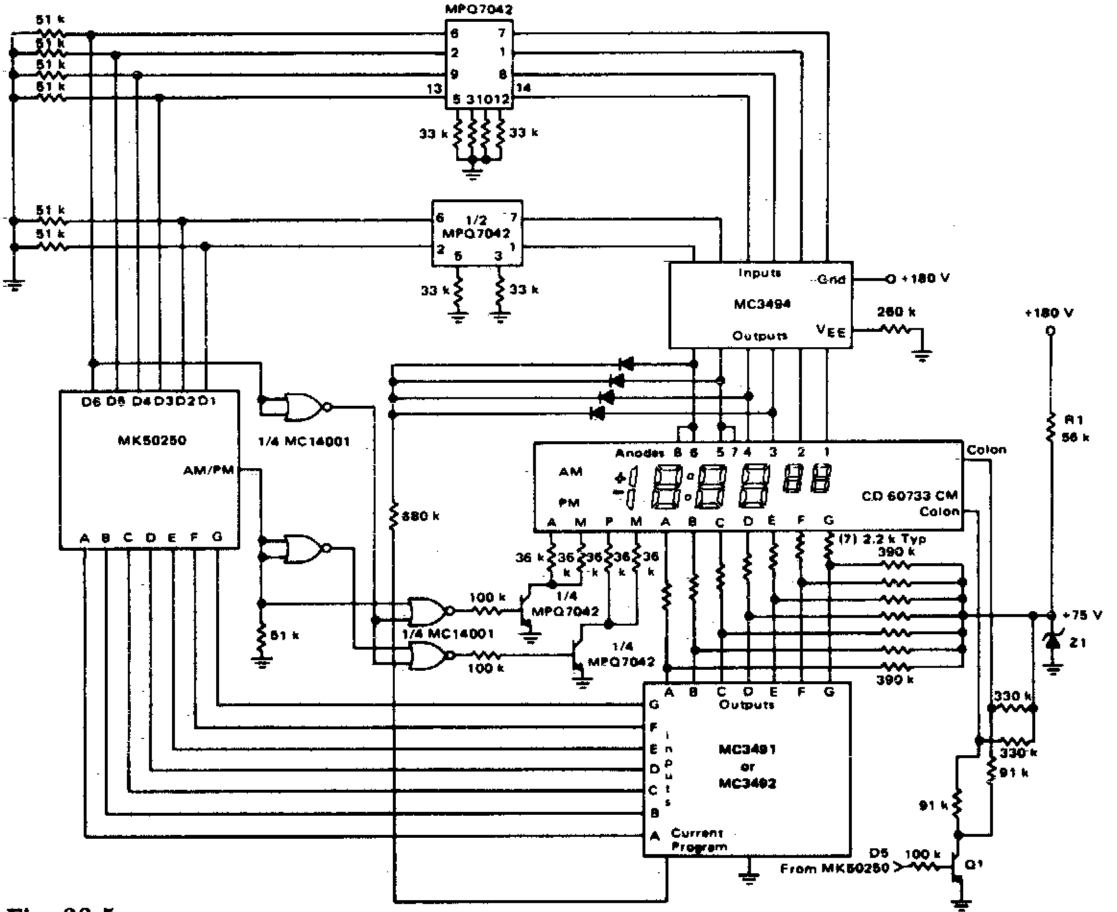

MPQ7042
51 k
33 k
1/2 MPQ7042
MC3494
Inputs
Gnd
Outputs
VEE
+180 V
260 k
R1
56 k
MK50250
D6 D5 D4 D3 D2 D1
AM/PM
A B C D E F G
1/4 MC14001
680 k
Anodes 8 6 5 7 4 3 2 1
AM
PM
CD 60733 CM
Colon
(7) 2.2 k Typ
36 k
390 k
100 k
1/4 MPQ7042
51 k
+75 V
Z1
330 k
91 k
MC3491 or MC3492
Outputs
Inputs
Current Program
D5
From MK50250
Q1

Fig. 28-5

PRECISION FREQUENCY COUNTER (~1 MHz MAXIMUM)

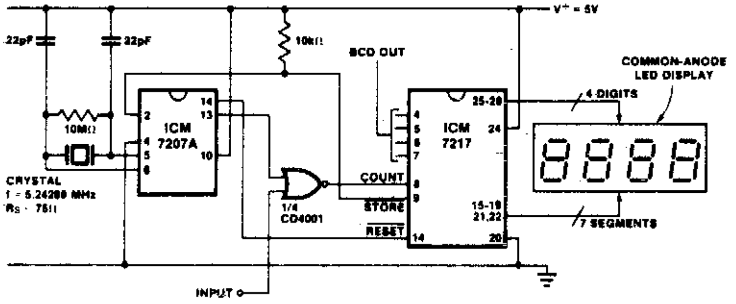
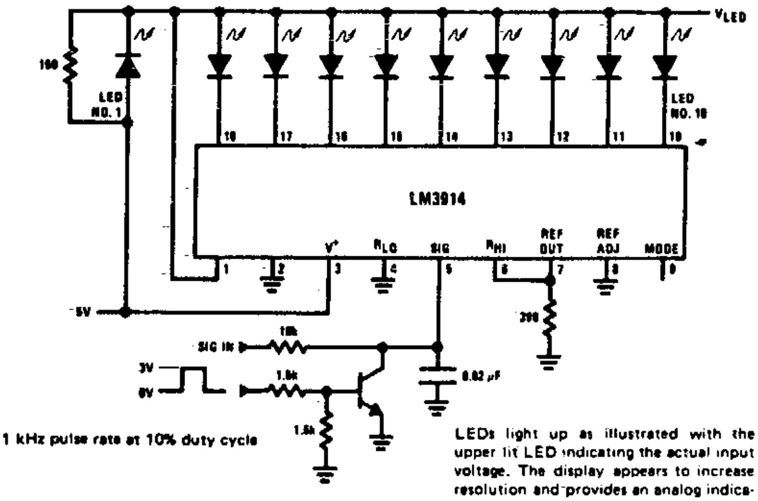

V+ = 5V
22pF
22pF
10MΩ
10kΩ
CRYSTAL
f = 5.24288 MHz
Rs ~ 75Ω
ICM 7207A
BCD OUT
1/4 CD4001
COUNT
STORE
RESET
ICM 7217
25-28
24
15-19 21,22
20
INPUT
4 DIGITS
7 SEGMENTS
COMMON-ANODE LED DISPLAY

Fig. 28-6

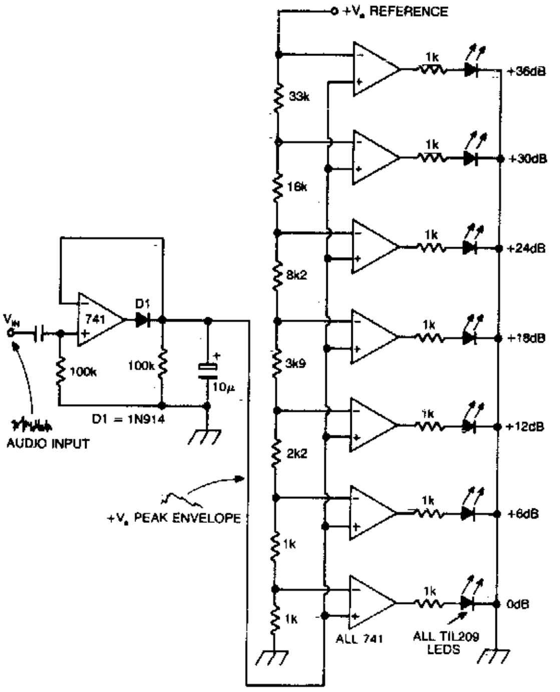
EXCLAMATION POINT DISPLAY


VLED
100
LED NO. 1
LED NO. 10
18
17
16
15
14
13
12
11
10
LM3914
V+
RLO
SIG
RHI
REF OUT
REF ADJ
MODE
1
2
3
4
5
6
7
8
9
5V
390
SIG IN
10k
3V
0V
1.0k
0.02 μF
1.5k
1 kHz pulse rate at 10% duty cycle

LEDs light up as illustrated with the upper lit LED indicating the actual input voltage. The display appears to increase resolution and provides an analog indication of overrange.

Fig. 28-7

LED BAR PEAK PROGRAM METER DISPLAY FOR AUDIO

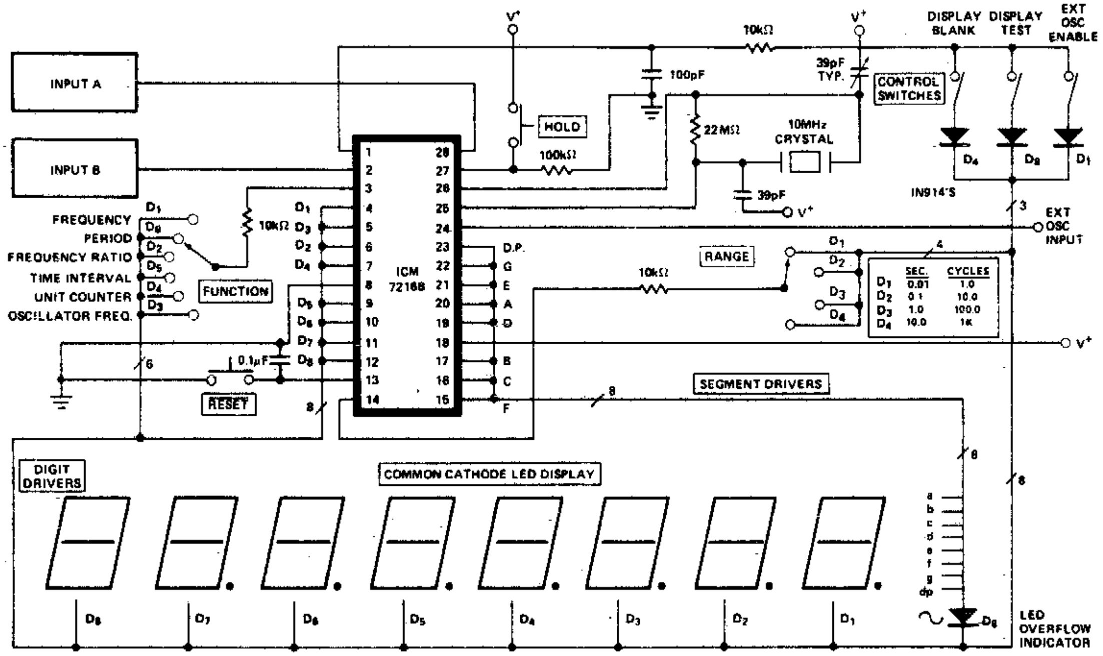

+Vs REFERENCE
33k
18k
8k2
3k9
2k2
1k
1k
1k
+36dB
+30dB
+24dB
+18dB
+12dB
+6dB
0dB
VIN
741
D1
100k
100k
10μ
D1 = 1N914
AUDIO INPUT
+Vs PEAK ENVELOPE
ALL 741
ALL TIL209 LEDS

Circuit Notes

A bar column of LEDs is arranged so that as the audio signal level increases, more LEDs in the column light up. The LEDs are arranged vertically in 6 dB steps. A fast response time and a one second decay time give an accurate response to transients and a low "flicker" decay characteristic. On each of the op amps inverting inputs is a dc reference voltage, which increases in 6 dB steps. All noninverting inputs are tied together and connected to the positive peak envelope of the audio signal. Thus, as this envelope exceeds a particular voltage reference, the op amp output goes high and the LED lights up. Also, all the LEDs below this are illuminated.

Fig. 28-8

10 MHz UNIVERSAL COUNTER



INPUT A
INPUT B
V+
10kΩ
100pF
39pF TYP.
CONTROL SWITCHES
DISPLAY BLANK
DISPLAY TEST
EXT OSC ENABLE
HOLD
100kΩ
22MΩ
10MHz CRYSTAL
39pF
IN914'S
EXT OSC INPUT
FREQUENCY
PERIOD
FREQUENCY RATIO
TIME INTERVAL
UNIT COUNTER
OSCILLATOR FREQ.
FUNCTION
10kΩ
ICM 7216B
D.P.
G
E
A
D
B
C
F
RANGE
SEC.
CYCLES
D1 0.01 1.0
D2 0.1 10.0
D3 1.0 100.0
D4 10.0 1K
0.1μF
RESET
SEGMENT DRIVERS
DIGIT DRIVERS
COMMON CATHODE LED DISPLAY
D8
D7
D6
D5
D4
D3
D2
D1
a b c d e f g dp
LED OVERFLOW INDICATOR

Fig. 28-9

Circuit Notes

This is a minimum component complete Universal Counter. It can use input frequencies up to 10 MHz at INPUT A and 2 MHz at INPUT B. If the signal at INPUT A has a very low duty cycle, it may be necessary to use a 74121 monostable multivibrator or similar circuit to stretch the input pulse width to be able to guarantee that it is at least 50 ns in duration.

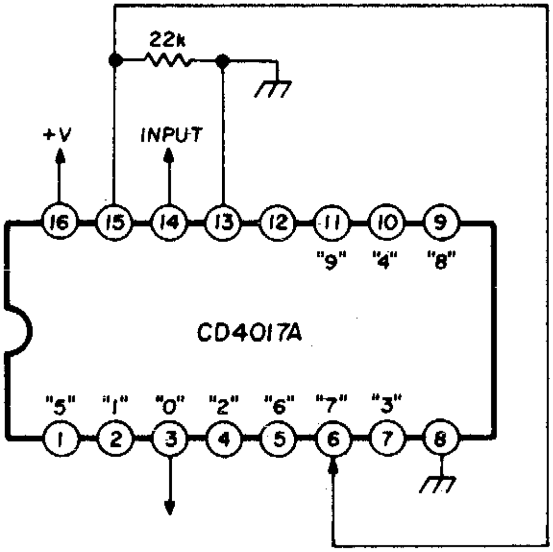
29

Dividers

The sources of the following circuits are contained in the Sources section beginning on page 730. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

CMOS Programmable Divide-by-N Counter
Frequency Divider Chain
Frequency Divider with Transient Free Output
Binary Divider Chain
Decade Frequency Divider

CMOS PROGRAMMABLE DIVIDE-BY-N COUNTER


22k
+V
INPUT
16
15
14
13
12
11
10
9
"9"
"4"
"8"
CD4017A
"5"
"1"
"0"
"2"
"6"
"7"
"3"
1
2
3
4
5
6
7
8

Circuit Notes

A single connection change permits division by any integer between 2 and 10. The RCA CD4017A Johnson decade counter is shown connected as a divide by 7 counter. The resistor is used to hold the reset line low. When the appropriate number is reached, that output and the reset line are driven high, resetting the counter. To divide by other integers, pin 15 should be connected to the desired output. For example, pin 1 for a divide by 5, or pin 7 for a divide by 3. The output of the divider appears on the 0 line.

Fig. 29-1

FREQUENCY DIVIDER CHAIN

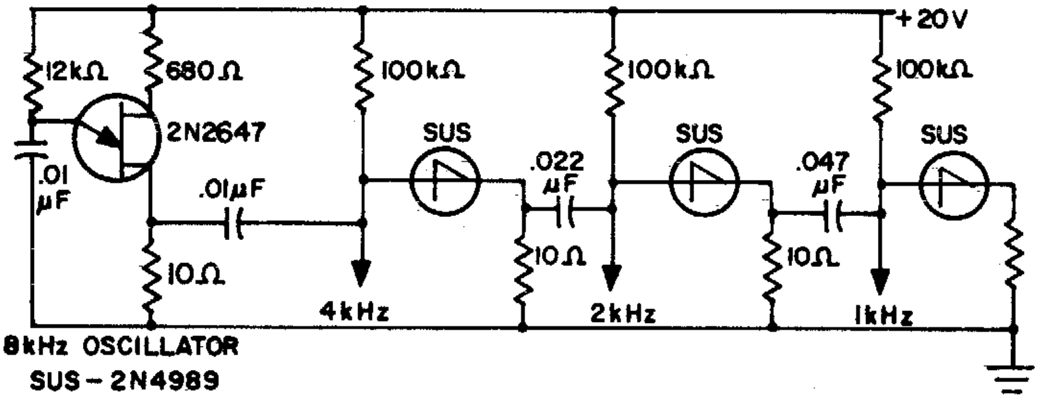
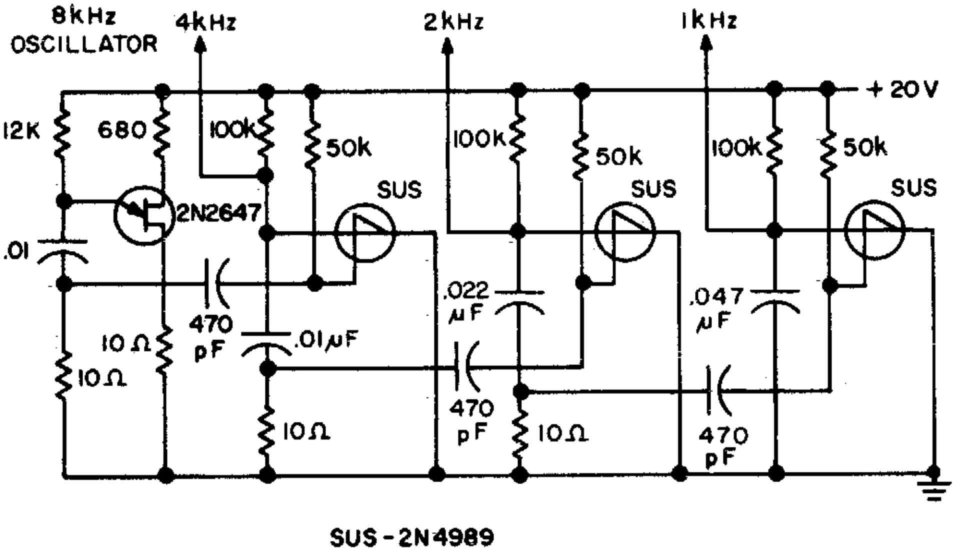

+20V
12kΩ
680Ω
2N2647
.01 μF
.01μF
10Ω
100kΩ
SUS
.022 μF
10Ω
4kHz
100kΩ
SUS
.047 μF
10Ω
2kHz
100kΩ
SUS
1kHz
8kHz OSCILLATOR
SUS – 2N4989

Fig. 29-2

Circuit Notes

Sawtooth output from each stage is one half frequency of preceding stage.

FREQUENCY DIVIDER WITH TRANSIENT FREE OUTPUT



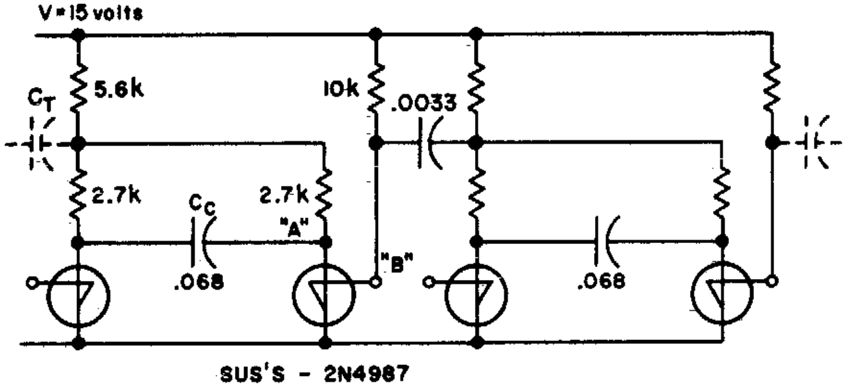
8kHz OSCILLATOR
4kHz
2kHz
1kHz
+20V
12K
680
100k
50k
SUS
2N2647
.01
470 pF
.01μF
10Ω
10Ω
10Ω
.022 μF
470 pF
10Ω
.047 μF
470 pF
SUS - 2N4989

Fig. 29-3

Circuit Notes

Spikes in the center of a sawtooth wave are eliminated in this circuit by triggering at gate.

BINARY DIVIDER CHAIN



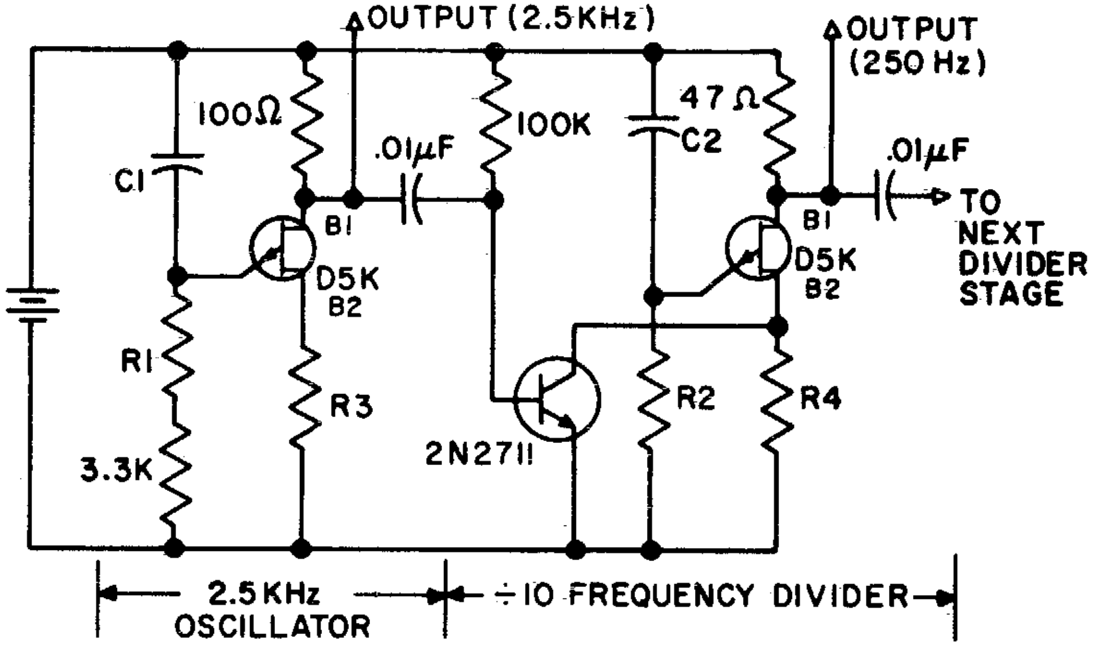
V=15 volts
C_T
5.6k
2.7k
C_C
2.7k
"A"
.068
10k
.0033
"B"
.068
SUS'S - 2N4987

Fig. 29-4

Circuit Notes

This circuit uses fewer components than transistor flip flops. Output at "B" gives a transient-free waveform.

DECADE FREQUENCY DIVIDER



OUTPUT (2.5KHz)
OUTPUT (250 Hz)
100Ω
100K
.01μF
47Ω
C2
.01μF
C1
B1
D5K
B2
TO NEXT DIVIDER STAGE
R1
R3
2N2711
R2
R4
3.3K
2.5KHz OSCILLATOR
÷10 FREQUENCY DIVIDER

Fig. 29-5

Circuit Notes

In the next stage, the product of R2 and C2 should be 10 × that of the preceding stage (±2%). R2 should be between 27K and 10 M.

C1 & C2—.0047 μF (±1 %)
R1—100K (±1%)
R2—1M (±1%)
R3—R4—1K (may need to be adjusted for variation of R_{BB} of UJT)

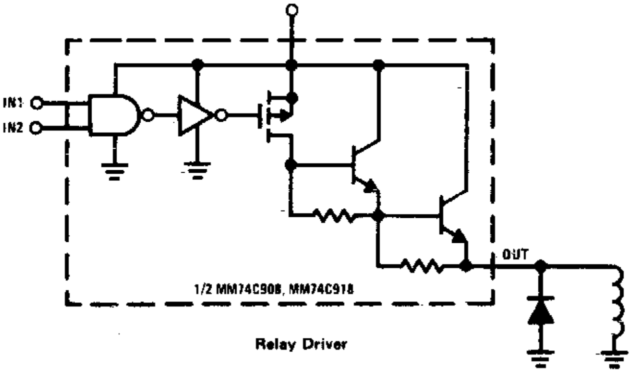
30

Drivers

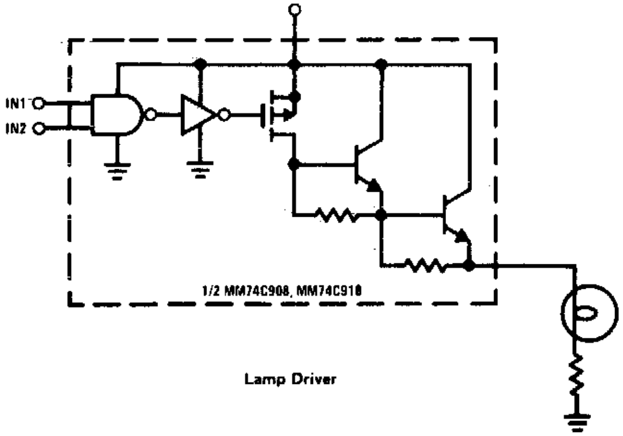
The sources of the following circuits are contained in the Sources section beginning on page 730. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

Driver Circuits
50 Ohm Driver
Line Driver
High Speed Laser Diode Driver
Capacitive Load Driver
Relay Driver
Relay Driver
BIFET Cable Driver
High Speed Line Driver for Multiplexers
High Impedance Meter Driver
CRT Deflection Yoke
CRT Yoke Driver
Solenoid Driver
Coaxial Cable Driver
High Speed Shield/Line Driver
Relay Driver with Strobe
Direct Dc Drive Interface of a Triac

DRIVER CIRCUITS

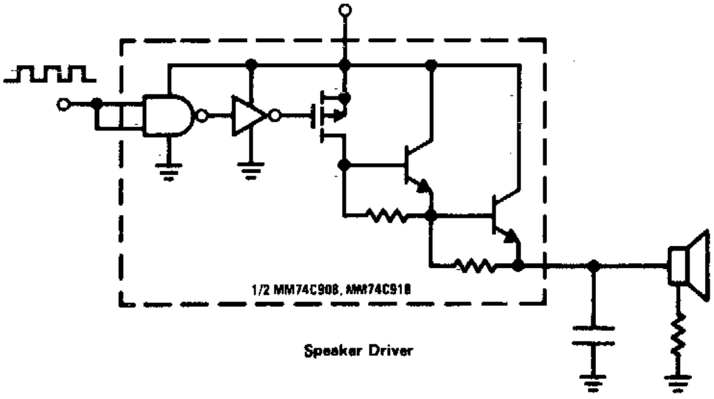

IN1
IN2
OUT
1/2 MM74C908, MM74C918

Relay Driver


IN1
IN2
1/2 MM74C908, MM74C918

Lamp Driver

Fig. 30-1

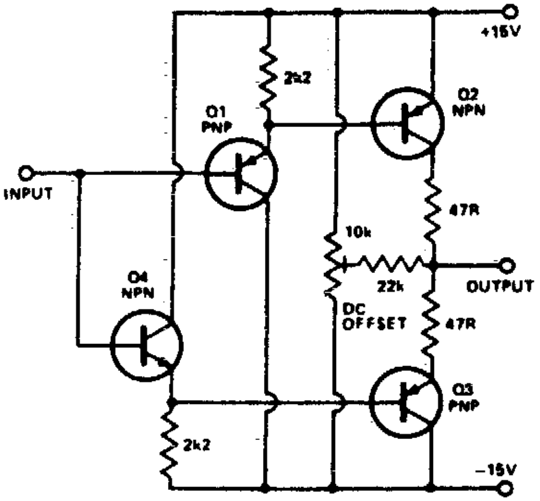

1/2 MM74C908, MM74C918

Speaker Driver

Circuit Notes

CMOS drivers for relays, lamps, speakers, etc., offers extremely low standby power. At $V_{CC} = 15$ V, power dissipation per package is typically 750 nW when the outputs are not drawing current. Thus, the drivers can be sitting out on line (a telephone line, for example) drawing essentially zero current until activated.

50 OHM DRIVER

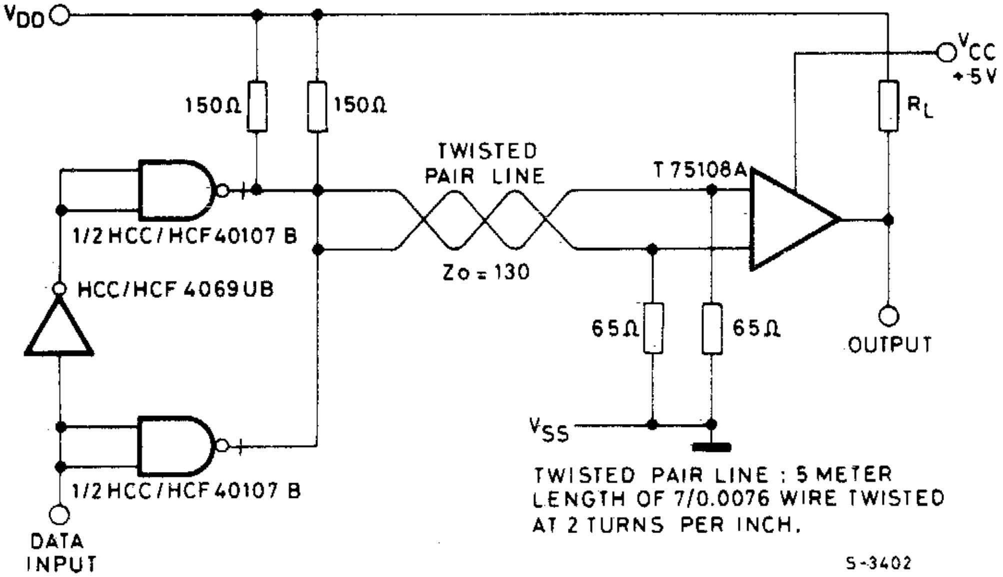

+15V
2k2
Q1
PNP
Q2
NPN
INPUT
47R
10k
22k
OUTPUT
DC
OFFSET
47R
Q4
NPN
Q3
PNP
2k2
-15V

Circuit Notes

To buffer a test generator to the outside world requires an amplifier with sufficient bandwidth and power handling capability. The circuit is a very simple unity gain buffer. It has a fairly high input impedance, a 50 ohm output impedance, a wide bandwidth, and high slew rate. The circuit is simply two pairs of emitter followers. The base emitter voltages of Q1 and Q2 cancel out, and so do those of Q3 and Q4. The preset is used to zero out any small dc offsets due to mismatching in the transistors.

Fig. 30-2

LINE DRIVER


VDD
150Ω
150Ω
TWISTED
PAIR LINE
T 75108A
VCC
+5V
RL
1/2 HCC/HCF40107 B
HCC/HCF 4069UB
Zo = 130
65Ω
65Ω
OUTPUT
VSS
1/2 HCC/HCF40107 B
DATA
INPUT
S-3402

TWISTED PAIR LINE : 5 METER LENGTH OF 7/0.0076 WIRE TWISTED AT 2 TURNS PER INCH.

Fig. 30-3

HIGH-SPEED LASER DIODE DRIVER

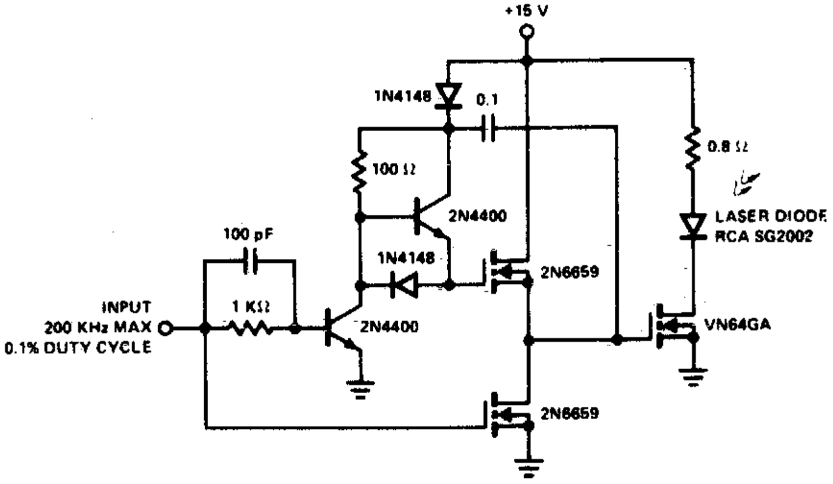
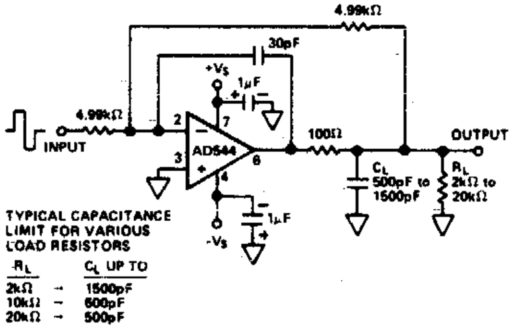

+15 V
1N4148
0.1
100 Ω
0.8 Ω
2N4400
LASER DIODE
RCA SG2002
100 pF
1N4148
2N6659
INPUT
200 KHz MAX
0.1% DUTY CYCLE
1 KΩ
2N4400
VN64GA
2N6659

Fig. 30-4

Circuit Notes

A faster driver can supply higher peak gate current to switch the VN64GA very quickly. The circuit uses a VMOS totempole stage to drive the high power switch.

CAPACITIVE LOAD DRIVER


4.99kΩ
30pF
+Vs
1μF
4.99kΩ
INPUT
2
7
3
AD544
6
4
100Ω
OUTPUT
CL
500pF to
1500pF
RL
2kΩ to
20kΩ
1μF
-Vs

TYPICAL CAPACITANCE LIMIT FOR VARIOUS LOAD RESISTORS

RL		CL UP TO
2kΩ	→	1500pF
10kΩ	→	600pF
20kΩ	→	500pF

Fig. 30-5

Circuit Notes

The circuit employs a 100 ohm isolation resistor which enables the amplifier to drive capacitive loads exceeding 500 pF; the resistor effectively isolates the high frequency feedback from the load and stabilizes the circuit. Low frequency feedback is returned to the amplifier summing junction via the low pass filter formed by the 100 ohm series resistor and the load capacitance. C_L.

RELAY DRIVER

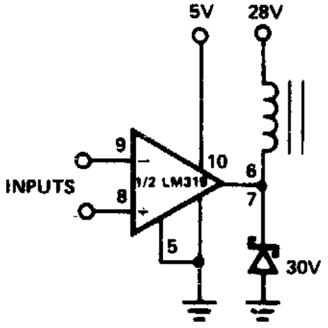

5V
28V
9
10
6
INPUTS
1/2 LM319
8
7
5
30V

Fig. 30-6

RELAY DRIVER

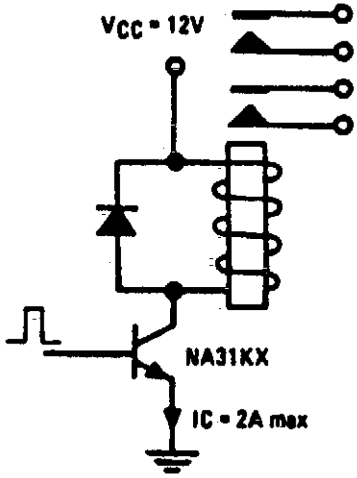
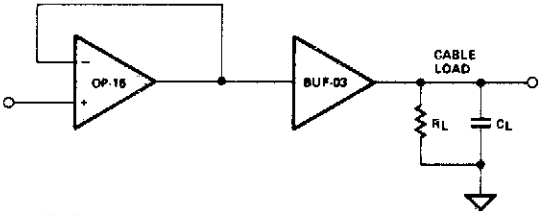

VCC = 12V
NA31KX
IC = 2A max

Fig. 30-7

BIFET CABLE DRIVER

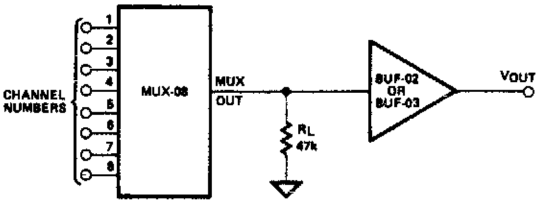

OP-15
BUF-03
CABLE LOAD
RL
CL

CAPACITIVE LOAD STABILITY OF BUF-03 MAKES IT AN IDEAL INTERFACE BETWEEN BIFET OP AMPS AND SHIELDED CABLES'

NOTE: TO MAINTAIN ACCURACY IN THE BUFFER $R_L > 1k\Omega$ IS RECOMMENDED.

Fig. 30-8

HIGH SPEED LINE DRIVER FOR MULTIPLEXERS


CHANNEL NUMBERS
1
2
3
4
5
6
7
8
MUX-08
MUX OUT
RL 47k
BUF-02 OR BUF-03
VOUT

NOTE 1: STRAY CAPACITANCE AT MULTIPLEXER OUTPUT NODE SHOULD BE MINIMIZED TO REDUCE CHANNEL-TO-CHANNEL CROSSTALK.

NOTE 2: A BUFFER WHOSE SLEW RATE IS TOO SMALL WILL INCREASE CHANNEL-TO-CHANNEL CROSSTALK.

Fig. 30-9

HIGH IMPEDANCE METER DRIVER

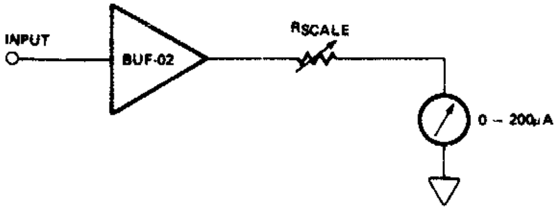

INPUT
BUF-02
RSCALE
0 – 200μA

Fig. 30-10

CRT DEFLECTION YOKE DRIVER

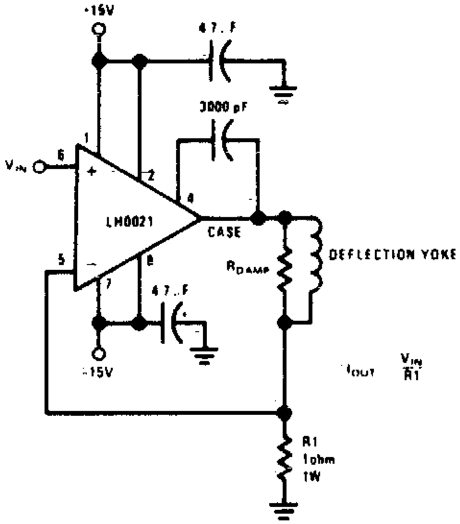
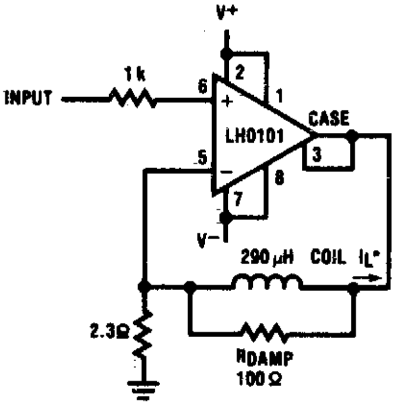

+15V
4.7μF
3000 pF
1
6
V_{IN}
+
2
4
LH0021
CASE
5
−
8
7
R_{DAMP}
DEFLECTION YOKE
4.7μF
−15V
$I_{OUT} = \frac{V_{IN}}{R1}$
R1
1ohm
1W

Fig. 30-11

CRT YOKE DRIVER


V+
INPUT
1k
6
+
2
1
LH0101
CASE
5
−
3
8
7
V−
290μH
COIL
I_L*
2.3Ω
R_{DAMP}
100Ω

*COIL CURRENT I_L MEASURED WITH TEKTRONIX CURRENT PROBE MODEL P6042

Fig. 30-12

Circuit Notes

A 500 mV peak-to-peak triangular waveform about ground is input to the amplifier, giving rise to a 100 mA peak current to the inductor.

SOLENOID DRIVER

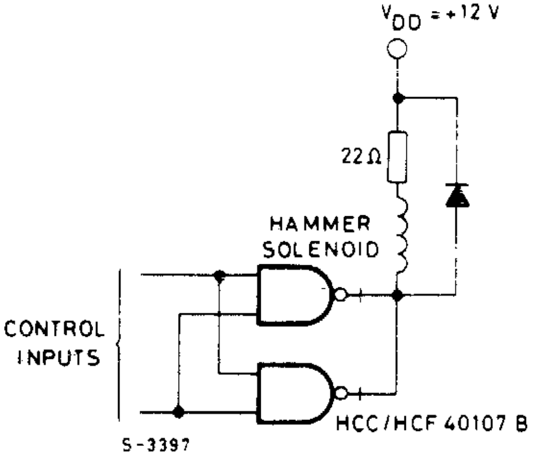
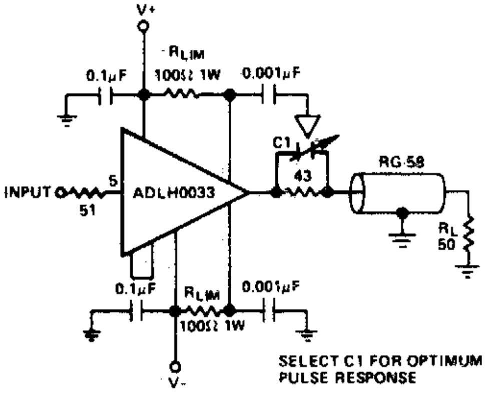

V_{DD} = +12 V
22Ω
HAMMER SOLENOID
CONTROL INPUTS
HCC/HCF 40107 B
S-3397

Fig. 30-13

COAXIAL CABLE DRIVER


V+
R_LIM
0.1μF
100Ω 1W
0.001μF
C1
43
RG-58
INPUT
51
5
ADLH0033
R_L
50
0.1μF
R_LIM
100Ω 1W
0.001μF
V-

SELECT C1 FOR OPTIMUM PULSE RESPONSE

Fig. 30-14

HIGH SPEED SHIELD/LINE DRIVER

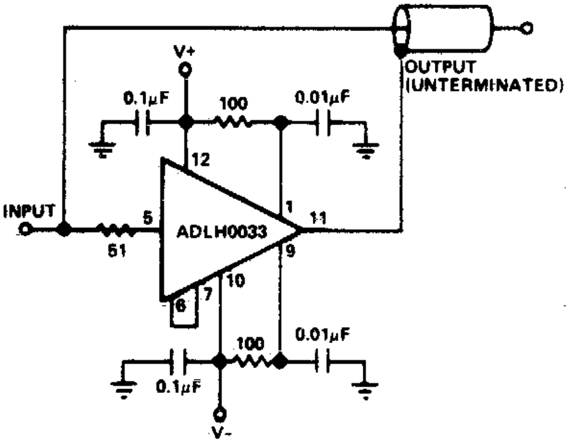
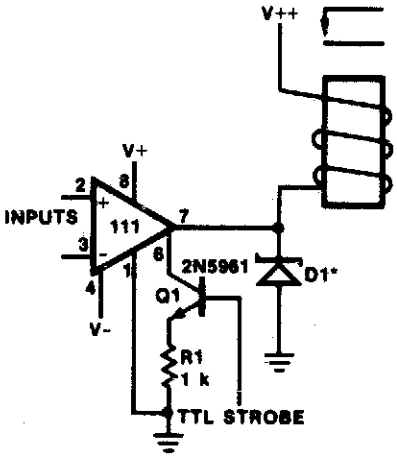

V+
0.1μF
100
0.01μF
OUTPUT
(UNTERMINATED)
12
INPUT
5
ADLH0033
1
11
9
51
8
7
10
100
0.01μF
0.1μF
V-

Fig. 30-15

RELAY DRIVER WITH STROBE


V++
V+
2
8
INPUTS
111
7
3
4
6
1
2N5961
D1*
Q1
V-
R1
1 k
TTL STROBE

*Absorbs inductive kickback of relay and protects IC from severe voltage transients on V++ line.

Fig. 30-16

DIRECT DC DRIVE INTERFACE OF A TRIAC

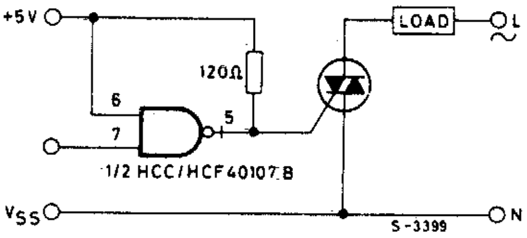

+5V
LOAD
L
~
120Ω
6
5
7
1/2 HCC/HCF40107B
V_SS
N
S-3399

Fig. 30-17

31

Fiber Optic Circuits

The sources of the following circuits are contained in the Sources section beginning on page 730. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

Fiber-Optics Half Duplex Information Link
Fiber-Optic Receiver, Very High Sensitivity, Low Speed, 3 nW
Fiber-Optic Link
Fiber-Optic Link Repeater
Fiber-Optic Receiver, High Sensitivity, 30 nW
Fiber-Optic Receiver, Low Sensitivity, 300 nW

FIBER-OPTICS HALF DUPLEX INFORMATION LINK

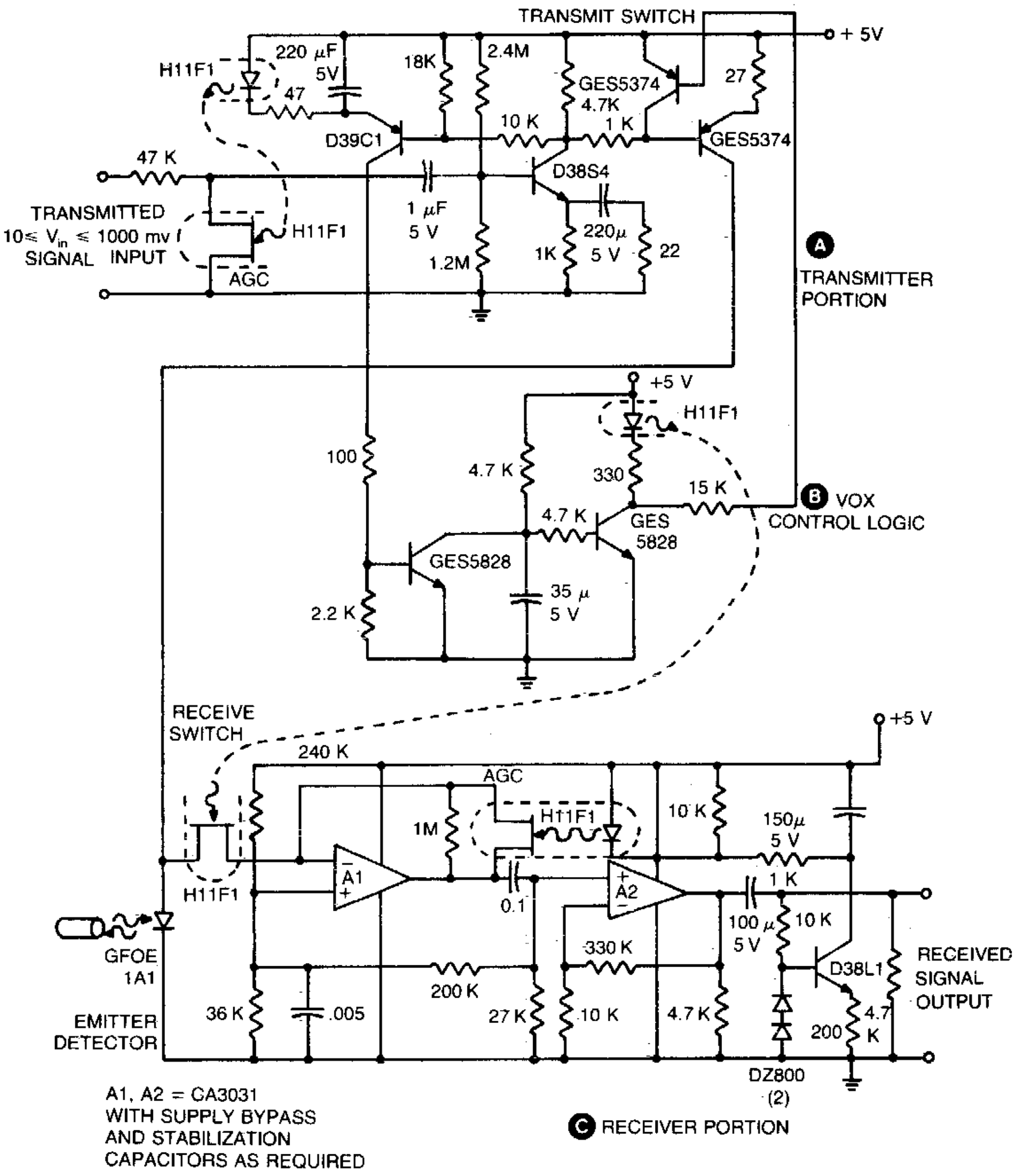

TRANSMIT SWITCH
+ 5V
H11F1
220 μF 5V
47
18K
2.4M
GES5374
4.7K
27
10 K
1 K
D39C1
GES5374
47 K
D38S4
TRANSMITTED 10≤ V_in ≤ 1000 mv SIGNAL INPUT
1 μF 5 V
220μ 5 V
H11F1
1K
22
1.2M
AGC
A TRANSMITTER PORTION
+5 V
H11F1
100
4.7 K
330
15 K
B VOX CONTROL LOGIC
4.7 K
GES 5828
GES5828
35 μ 5 V
2.2 K
RECEIVE SWITCH
+5 V
240 K
AGC
1M
H11F1
10 K
150μ 5 V
1 K
A1
A2
0.1
H11F1
GFOE 1A1
EMITTER DETECTOR
100 μ 5 V
10 K
330 K
D38L1
RECEIVED SIGNAL OUTPUT
36 K
.005
200 K
27 K
10 K
4.7 K
200
4.7 K
DZ800 (2)
C RECEIVER PORTION
A1, A2 = CA3031 WITH SUPPLY BYPASS AND STABILIZATION CAPACITORS AS REQUIRED

Fig. 31-1

FIBER-OPTIC RECEIVER, VERY HIGH SENSITIVITY, LOW SPEED, 3nW

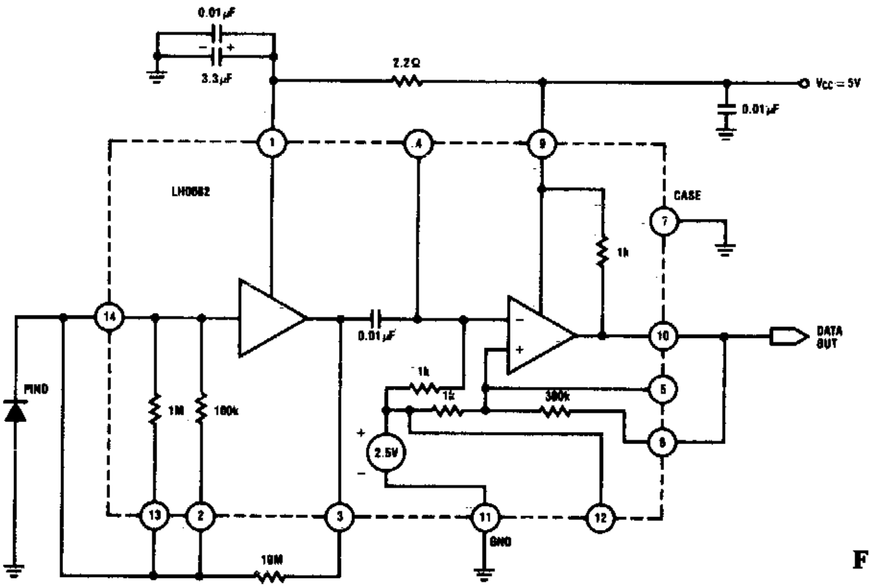

0.01µF
3.3µF
2.2Ω
VCC = 5V
0.01µF
LH0082
CASE
1k
PIND
1M
100k
0.01µF
1k
1k
300k
2.5V
GND
10M
DATA OUT
1
4
9
7
14
10
5
6
13
2
3
11
12

Fig. 31-2

FIBER-OPTIC LINK

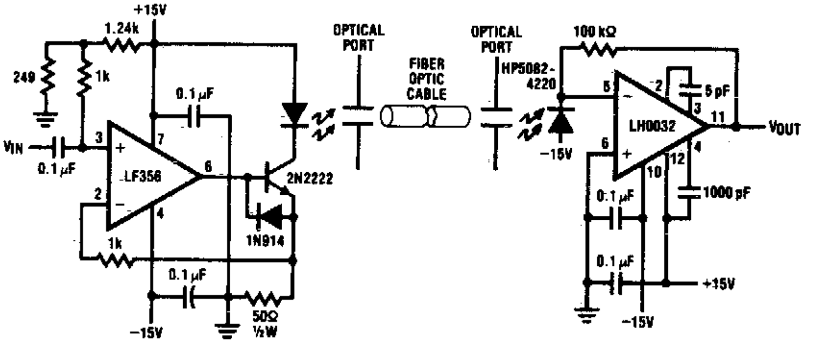
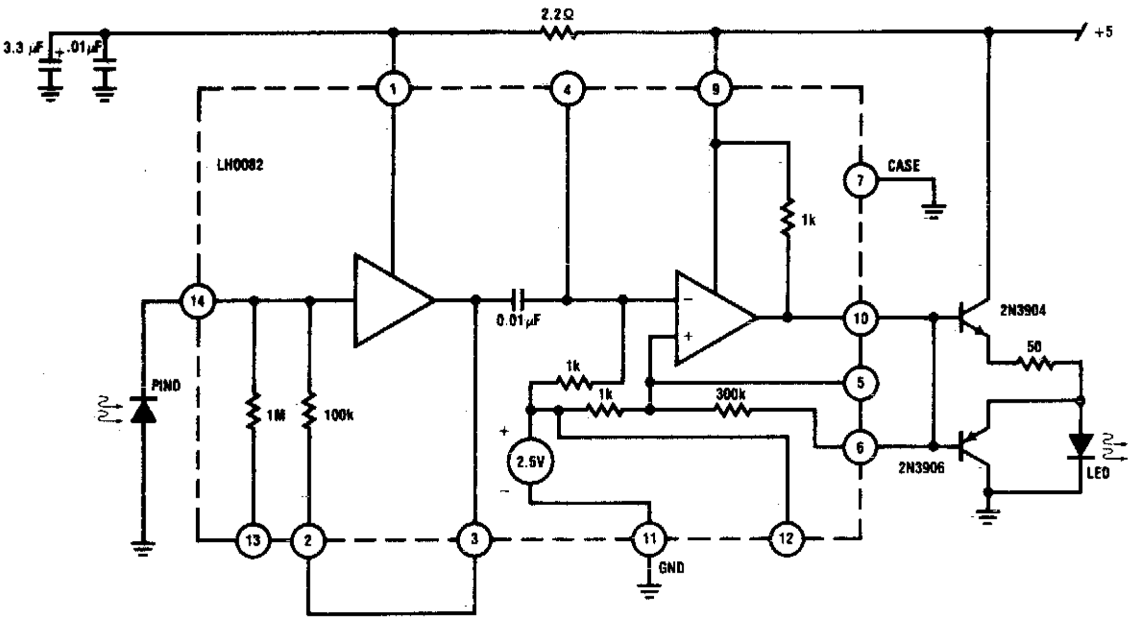

+15V
1.24k
249
1k
0.1µF
VIN
0.1µF
LF356
2N2222
1N914
1k
0.1µF
50Ω ½W
−15V
OPTICAL PORT
FIBER OPTIC CABLE
OPTICAL PORT
HP5082-4220
100 kΩ
−15V
LH0032
5pF
1000pF
VOUT
0.1µF
0.1µF
+15V
−15V

Fig. 31-3

Circuit Notes

Fiber Optic applications require analog drivers and receivers operating in the megahertz region. This complete analog transmission system is suitable for optical communication applications up to 3.5 MHz. The transmitter LED is normally biased at 50 mA operating current. The input is capacitively coupled and ranges from 0 to 5 V, modulating the LED current from 0 to 100 mA. The receiver circuit is configured as a transimpedance amplifier. The photodiode with 0.5 amp per watt responsivity generates a 50 mV signal at the receiver output for 1 μW of light input.

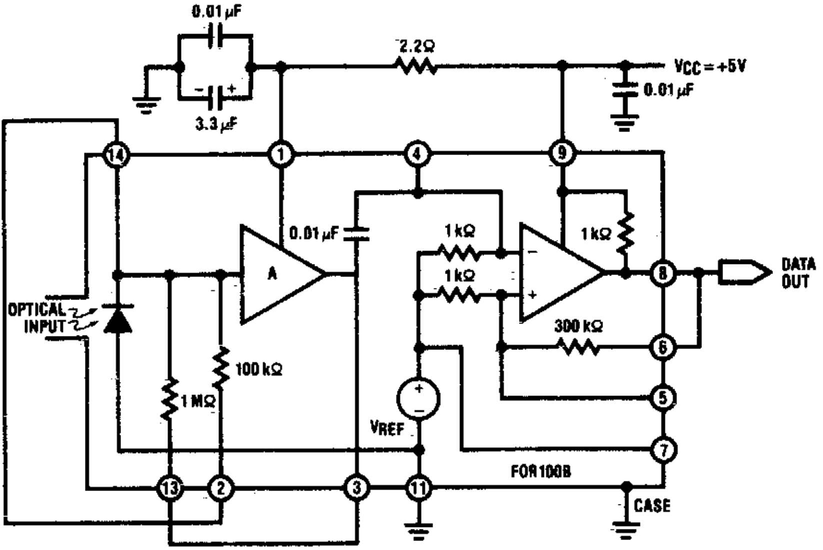
FIBER-OPTIC LINK REPEATER



3.3 μF
+
.01μF
2.2Ω
+5
LH0082
1
4
9
CASE
7
1k
14
0.01 μF
−
+
10
2N3904
50
PIND
1M
100k
1k
1k
300k
5
6
2N3906
LED
+
2.5V
−
13
2
3
11
12
GND

Fig. 31-4

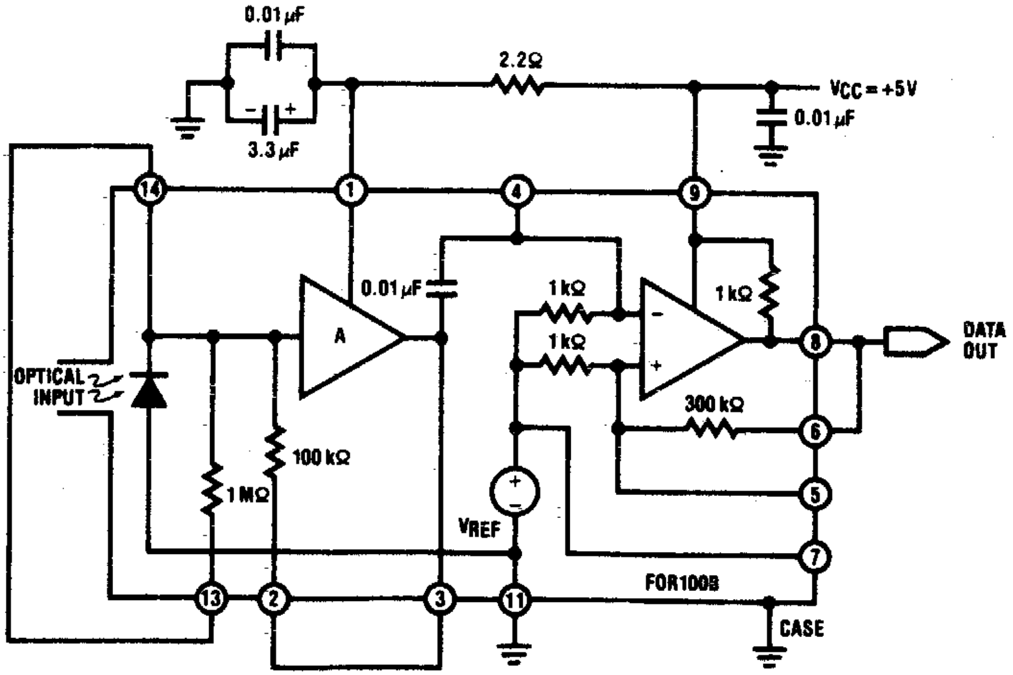
FIBER-OPTIC RECEIVER, HIGH SENSITIVITY, 30nW



0.01 μF
2.2Ω
VCC = +5V
0.01 μF
−
+
3.3 μF
14
1
4
9
A
0.01 μF
1 kΩ
1 kΩ
1 kΩ
−
+
8
DATA OUT
300 kΩ
6
5
7
OPTICAL INPUT
1 MΩ
100 kΩ
+
−
VREF
FOR100B
13
2
3
11
CASE

Fig. 31-5

FIBER-OPTIC RECEIVER, LOW SENSITIVITY, 2 μW



0.01 μF
2.2 Ω
VCC = +5V
0.01 μF
3.3 μF
14
1
4
9
0.01 μF
1 kΩ
1 kΩ
1 kΩ
A
OPTICAL INPUT
8
DATA OUT
300 kΩ
6
100 kΩ
1 MΩ
5
VREF
7
FOR100B
13
2
3
11
CASE

Fig. 31-6

32

Field Strength Meters

The sources of the following circuits are contained in the Sources section beginning on page 730. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

Low Cost Microwave Field Strength Meter
Sensitive Field-Strength Meter
Adjustable Sensitivity Field-Strength Indicator
Field Strength Meter – 1.5 to 150 MHz
Simple Field Strength Meter
Untuned Field Strength Meter
Tuned Field Strength Meter
VOM Field Strength Meter

LOW COST MICROWAVE FIELD STRENGTH METER

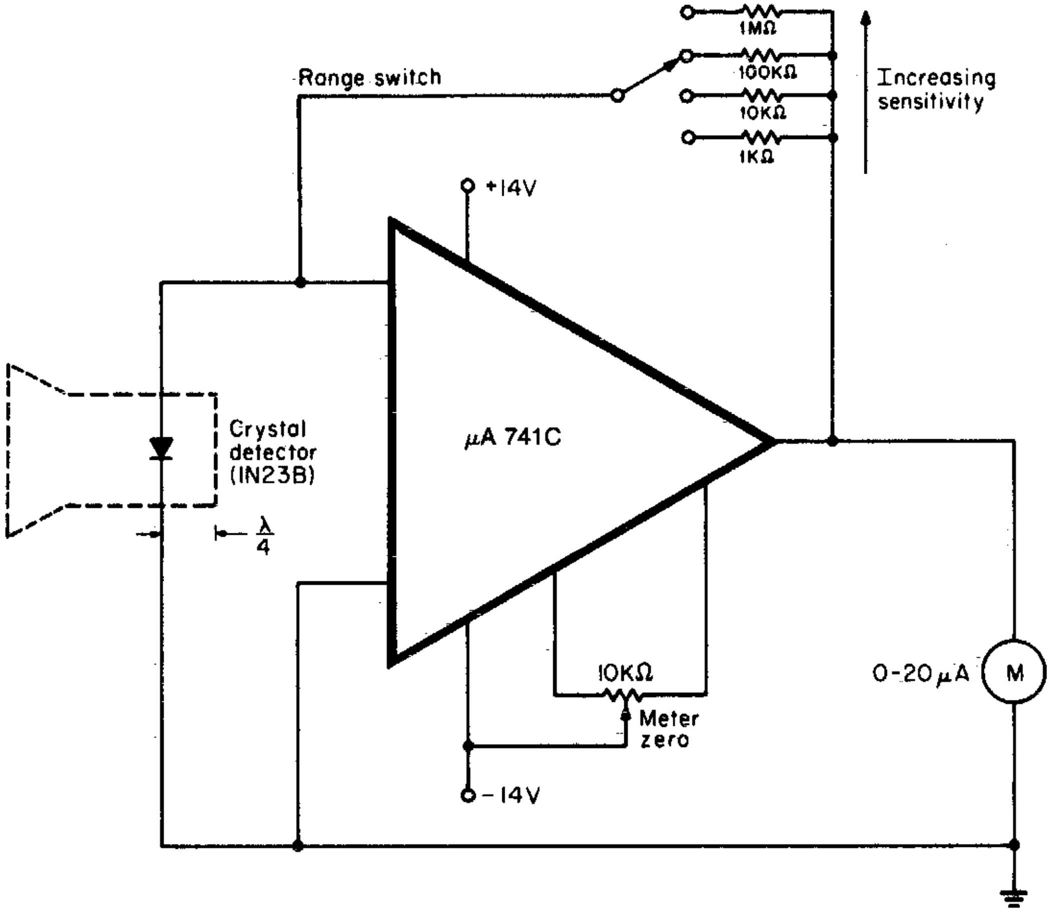

1MΩ
100KΩ
10KΩ
1KΩ
Range switch
Increasing sensitivity
+14V
μA 741C
Crystal detector (IN23B)
λ/4
10KΩ
Meter zero
-14V
0-20μA
M

Fig. 32-1

Circuit Notes

When operating, a waveguide directs energy onto a crystal detector. The diode shown is for X-band operation. The waveguide is a 1½ inch piece of plastic tubing with the ends flared. The plastic is coated with an electroless copper solution to provide a conducting surface. The dimensions are not critical. For calibrated readings, the meter is placed in a known field or else compared to a calibrated meter. To operate the meter, point it away from the signal. Switch the meter to the desired range, and adjust the zero control for a 0 reading. Then point the waveguide at the signal, and read field strength directly.

SENSITIVE FIELD-STRENGTH METER

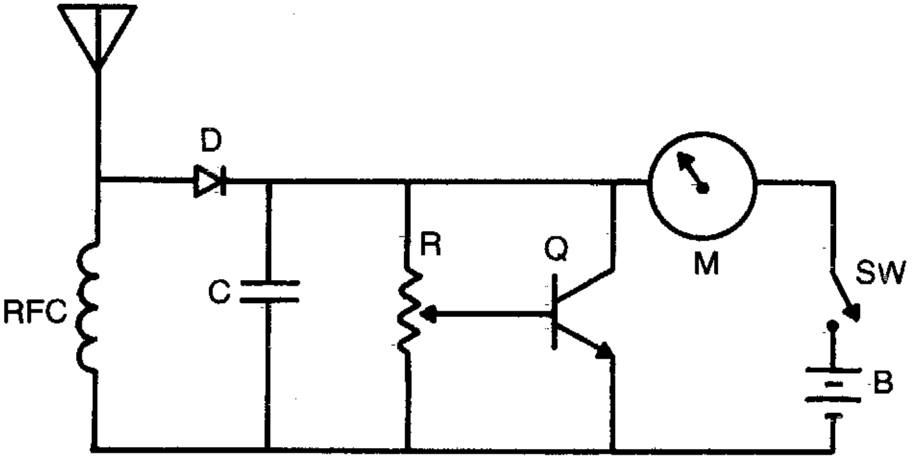
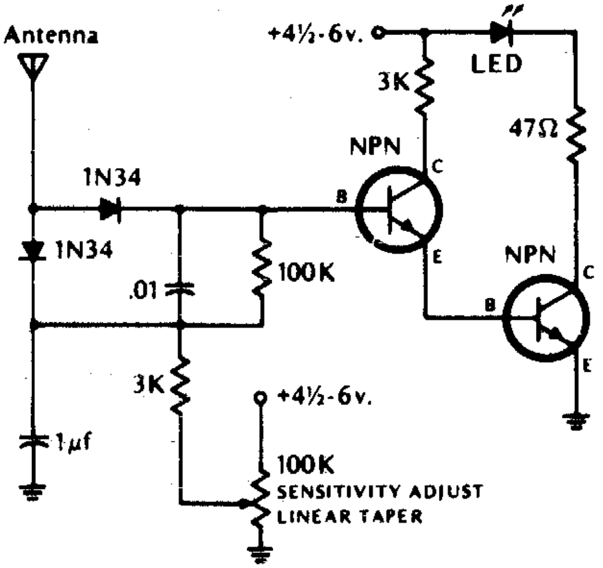

D
R
Q
M
SW
RFC
C
B

Fig. 32-2

Circuit Notes

Increased sensitivity gives field strength reading from low power transmitters. Operating range 3-30 MHz. To operate, adjust R for ⅓ to ½ scale reading. RFC = 2.5 mH choke, C = 1,000 pF, R = 50 K pot, M = 0 – 1 mA, D = 1N34 or 1N60 (Germanium), Q = NPN (RCASK3020, 2N3904 or equivalent).

ADJUSTABLE-SENSITIVITY FIELD-STRENGTH INDICATOR


Antenna
+4½-6v.
3K
LED
47Ω
NPN
C
B
E
1N34
1N34
.01
100K
NPN
C
B
E
3K
+4½-6v.
1µf
100K
SENSITIVITY ADJUST
LINEAR TAPER

Circuit Notes

The LED lights if the rf field is higher than the pre-set field strength level. Diodes should be germanium. Transistors (NPN) = 2N2222, 2N3393, 2N3904 or equivalent.

Fig. 32-3

FIELD STRENGTH METER – 1.5 to 150 MHz

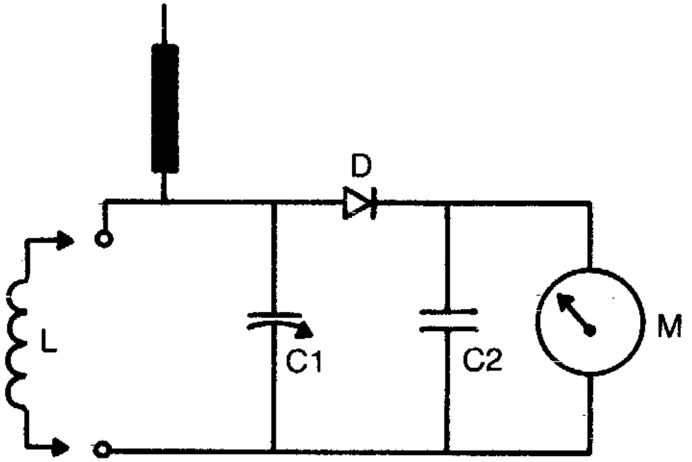
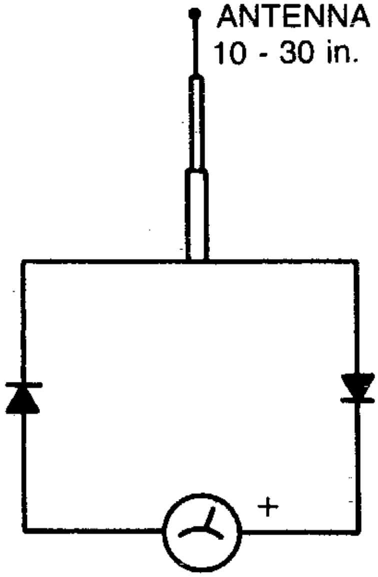


Fig. 32-4

Circuit Notes

The tuning range is determined by coil (L) dimensions and setting of C1. Coils can be plugged in for multirange use or soldered in place if only limited frequency range is of interest. C1 = 36 pF variable, C2 = .0047 disc, D = 1N60 (germanium) and M = 0–1 mA meter. For increased sensitivity, use 50 μA meter.

SIMPLE FIELD STRENGTH METER

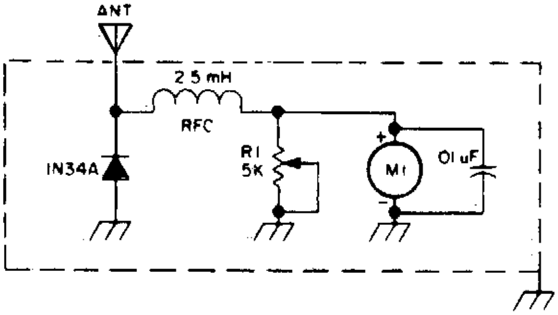


Circuit Notes

The circuit is frequency selective. It has been used from 2 meters through 160 meters. The telescoping antenna may be adjusted to its shortest length when working at 2 meters to keep the needle on the scale. Meter should be a 100 microamp to a 500 microamp movement. The diodes are germanium type, such as 1N34, etc. Silicon diodes will also work, but they are a bit less sensitive.

Fig. 32-5

UNTUNED FIELD STRENGTH METER

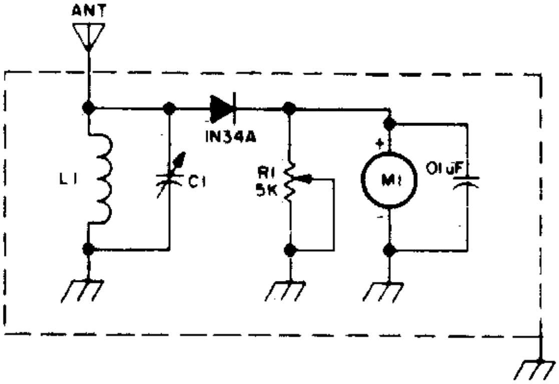

ANT
2.5 mH
RFC
IN34A
R1
5K
+
M1
-
.01 uF

Circuit Notes

Sensitivity is controlled by R1 and sensitivity of Meter M1.

Fig. 32-6

TUNED FIELD STRENGTH METER


ANT
IN34A
L1
C1
R1
5K
+
M1
-
.01 uF

Circuit Notes

Resonant combination of L1 and C1 are selected to cover frequencies desired.

Fig. 32-7

VOM FIELD STRENGTH METER

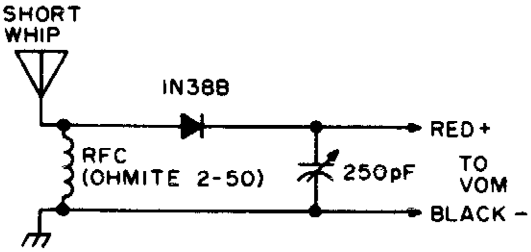

SHORT
WHIP
IN38B
RFC
(OHMITE 2-50)
250pF
RED +
TO
VOM
BLACK −

Fig. 32-8

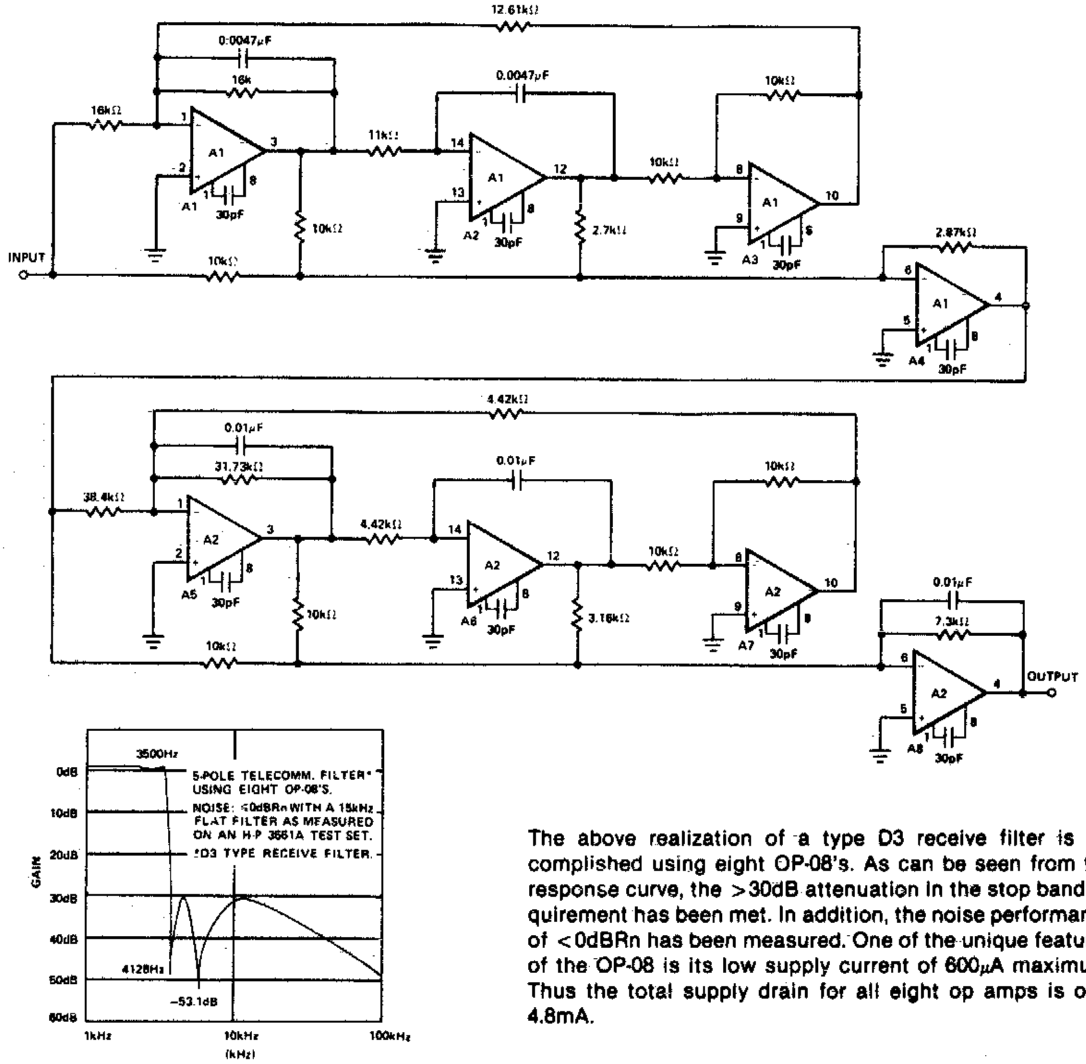
33

Filters

The sources of the following circuits are contained in the Sources section beginning on page 730. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

Five-Pole Active Filter
Digitally Tuned Low Power Active Filter
10 kHz Sallen-Key Low-Pass Filter
Fourth Order High-Pass Butterworth Filter
Tunable Notch Filter to Suppress Hum
Three Amplifier Notch Filter (or Elliptical Filter Building Block)
Selectable Bandwidth Notch Filter
4.5 MHz Notch Filter
High Q Notch Filter
Rejection Filter
Notch Filter Using the μA 4136 as a Gyrator
1 kHz Bandpass Active Filter
Bandpass Active Filter with 60 dB Gain
Multiple Feedback Bandpass Filter
Biquad RC Active Bandpass Filter
400 Hz Low-Pass Butterworth Active Filter
Variable Bandwidth Bandpass Active Filter
Low-Pass Filter
High Q Bandpass Filter
MFB Bandpass Filter for Multichannel Tone Decoder
Sallen-Key Second Order Low-Pass Filter
Three Amplifier Active Filter
Bandpass State Variable Filter
Universal State Variable Filter
500 Hz Sallen-Key Bandpass Filter
Filter Networks
Equal Component Sallen-Key Low-Pass Filter
Biquad Filter
Second Order State Variable Fitter (1 kHz, Q = 10)
Biquad Filter
Tunable Active Filter
Active RC Filter for Frequencies up to 150 kHz
Pole Active Low-Pass Filter (Butterworth Maximally Flat Response)
Speech Filter (300 Hz .3 kHz Bandpass)
0.1 Hz to 10 Hz Bandpass Filter
High-Pass Active Filter
Second Order High-Pass Active Filter
High Pass Filter (High Frequency)
160 Hz Bandpass Filter
Multiple Feedback Bandpass Filter (1.0 kHz)
20 kHz Bandpass Active Filter
Rumble Filter Using LM387
Scratch Filter Using LM287

FIVE-POLE ACTIVE FILTER



12.61kΩ
0.0047μF
16k
16kΩ
11kΩ
0.0047μF
10kΩ
10kΩ
INPUT
10kΩ
10kΩ
2.7kΩ
2.87kΩ
A1
A2
A3
A4
30pF
4.42kΩ
0.01μF
31.73kΩ
38.4kΩ
4.42kΩ
0.01μF
10kΩ
10kΩ
10kΩ
3.16kΩ
0.01μF
7.3kΩ
A5
A6
A7
A8
OUTPUT
3500Hz
5-POLE TELECOMM. FILTER*
USING EIGHT OP-08'S.
NOISE: ≤0dBRn WITH A 15kHz
FLAT FILTER AS MEASURED
ON AN H-P 3561A TEST SET.
*D3 TYPE RECEIVE FILTER.
0dB
10dB
20dB
30dB
40dB
50dB
60dB
GAIN
4128Hz
−53.1dB
1kHz
10kHz
100kHz
(kHz)

The above realization of a type D3 receive filter is accomplished using eight OP-08's. As can be seen from the response curve, the >30dB attenuation in the stop band requirement has been met. In addition, the noise performance of <0dBRn has been measured. One of the unique features of the OP-08 is its low supply current of 600μA maximum. Thus the total supply drain for all eight op amps is only 4.8mA.

Fig. 33-1

DIGITALLY TUNED LOW POWER ACTIVE FILTER

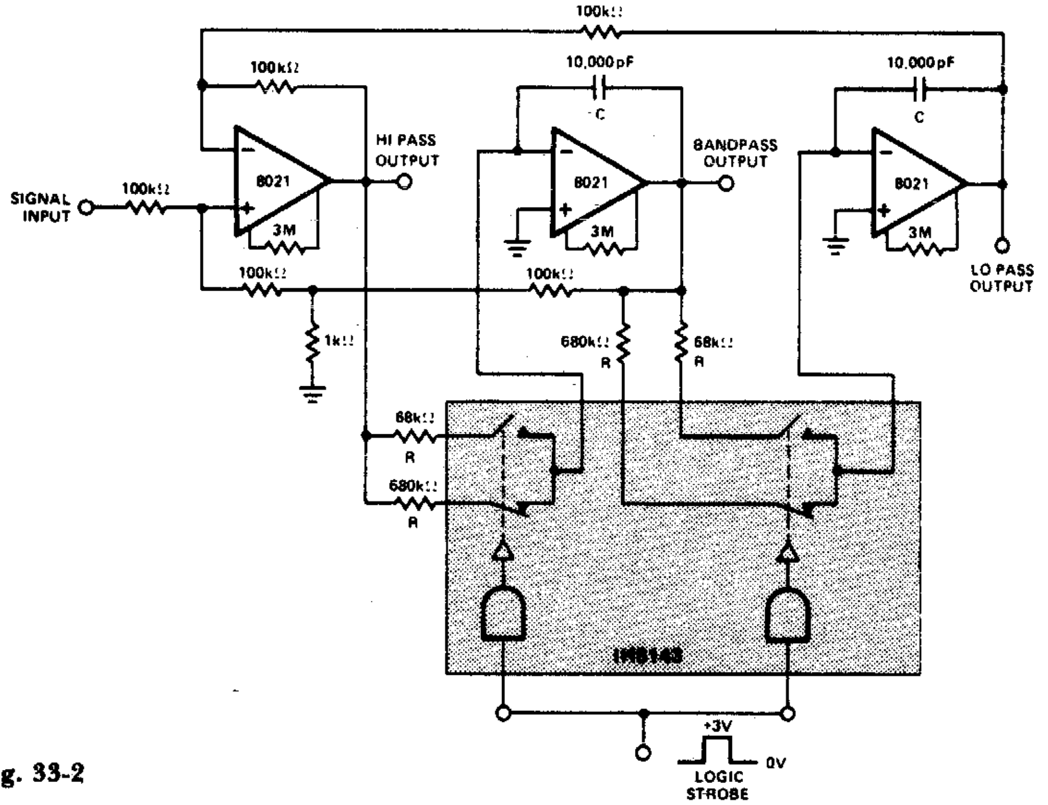

100kΩ
100kΩ
10,000pF
C
10,000pF
C
HI PASS OUTPUT
BANDPASS OUTPUT
SIGNAL INPUT
100kΩ
8021
8021
8021
3M
3M
3M
LO PASS OUTPUT
100kΩ
100kΩ
1kΩ
680kΩ R
68kΩ R
68kΩ R
680kΩ R
IH5143
+3V
0V
LOGIC STROBE

Fig. 33-2

Circuit Notes

Constant gain, constant Q, variable frequency filter which provides simultaneous low-pass, bandpass, and high-pass outputs. With the component values shown, center frequency will be 235 Hz and 23.5 Hz for high and low logic inputs respectively, Q = 100, and gain = 100.

$$f_n = \text{center frequency} = \frac{1}{2\pi RC}$$

10 kHz SALLEN-KEY LOW-PASS FILTER

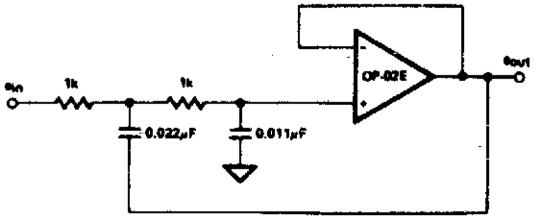
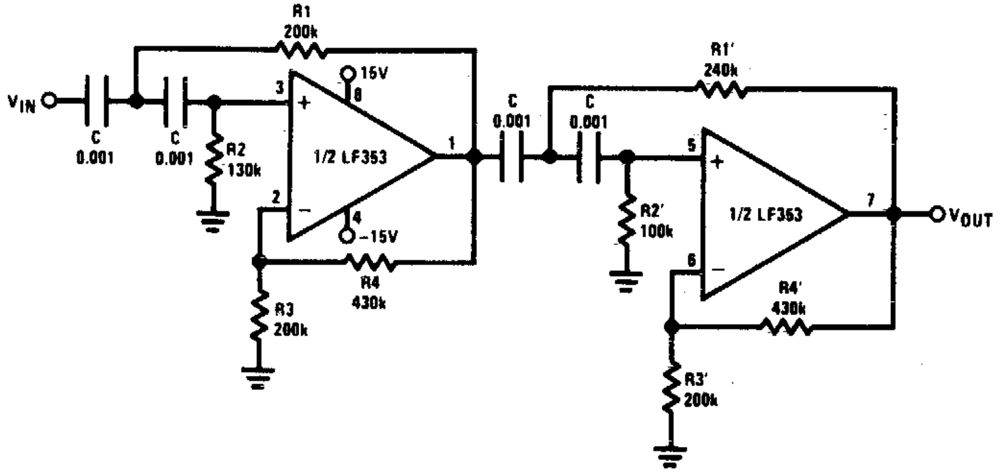

e_{in}
1k
1k
0.022μF
0.011μF
OP-02E
e_{out}

Fig. 33-3

FOURTH ORDER HIGH-PASS BUTTERWORTH FILTER

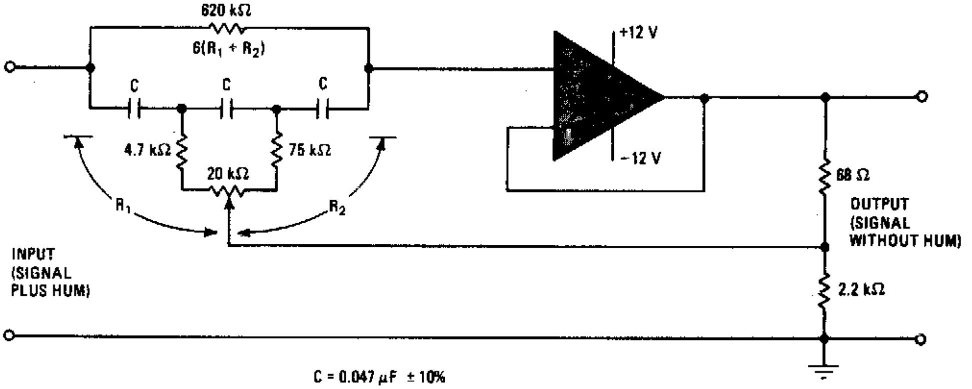


R1
200k
15V
8
R1'
240k
V_IN
3
+
C
0.001
C
0.001
R2
130k
1/2 LF353
1
C
0.001
C
0.001
5
+
2
−
4
−15V
R4
430k
R3
200k
R2'
100k
1/2 LF353
7
V_OUT
6
−
R4'
430k
R3'
200k

- Corner frequency $(f_C) = \sqrt{\dfrac{1}{R1R2C^2}} \cdot \dfrac{1}{2\pi} = \sqrt{\dfrac{1}{R1'R2'C^2}} \cdot \dfrac{1}{2\pi}$
- Passband gain $(H_O) = (1 + R4/R3)(1 + R4'/R3')$
- First stage Q = 1.31
- Second stage Q = 0.541
- Circuit shown uses closest 5% tolerance resistor values for a filter with a corner frequency of 1 kHz and a passband gain of 10

Fig. 33-4

TUNABLE NOTCH FILTER TO SUPPRESS HUM



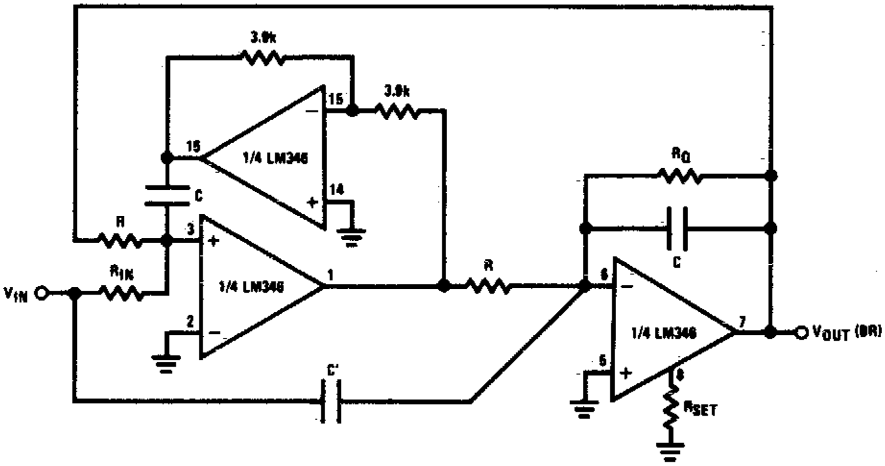
620 kΩ
6(R₁ + R₂)
C
C
C
+12 V
−12 V
4.7 kΩ
75 kΩ
20 kΩ
R₁
R₂
68 Ω
OUTPUT
(SIGNAL
WITHOUT HUM)
INPUT
(SIGNAL
PLUS HUM)
2.2 kΩ
C = 0.047 μF ± 10%

Fig. 33-5

Circuit Notes

This narrow-stop-band filter can be tuned by the pot to place the notch at any frequency from 45 to 90 Hz. It attenuates power-line hum or other unwanted signals by at least 30 dB. Because the circuit uses wide-tolerance parts, it is inexpensive to build.

THREE-AMPLIFIER NOTCH FILTER (OR ELLIPTIC FILTER BUILDING BLOCK)



3.9k
3.9k
15
16
14
1/4 LM346
C
R
3
1/4 LM346
1
R_{IN}
V_{IN}
2
C'
R
R_Q
C
6
1/4 LM346
7
V_{OUT} (BR)
5
8
R_{SET}

Circuit Synthesis Equations

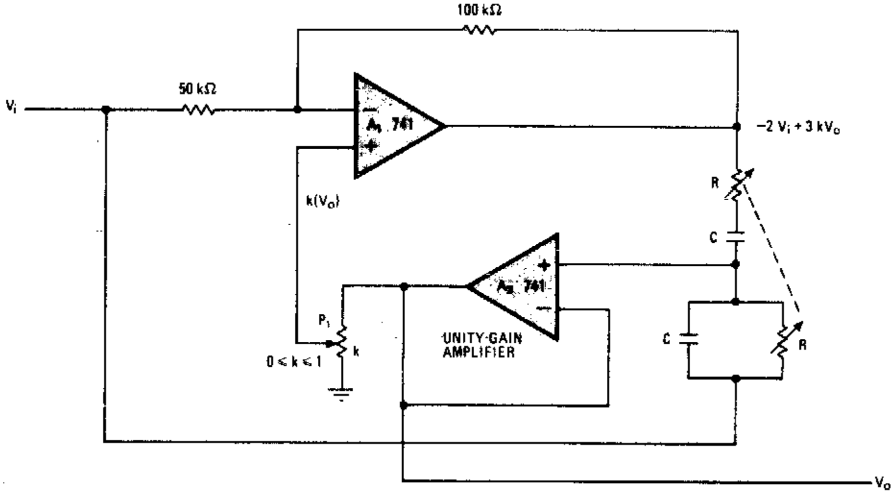
$$R \times C = \frac{0.159}{f_o}\ ; R_Q = Q_o \times R;\ R_{IN} = \frac{0.159 \times f_o}{C' \times f^2_{notch}}$$

- For nothing but a notch output: $R_{IN} = R$, $C' = C$.

$$H_{o(BR)}\Big|_{f << f_{notch}} = \frac{R}{R_{IN}}\quad H_{o(BR)}\Big|_{f >> f_{notch}} = \frac{C'}{C}$$

Fig. 33-6

SELECTABLE BANDWIDTH NOTCH FILTER



100 kΩ
50 kΩ
V_i
A_1 741
$-2V_i + 3kV_o$
R
k(V_o)
C
A_2 741
UNITY-GAIN AMPLIFIER
P_1
k
$0 \leq k \leq 1$
C
R
V_o

Fig. 33-7

Circuit Notes

This notch filter, which operates at up to 200 kHz, uses a modified Wien bridge to select bandwidth over which frequencies are rejected. RC components determine filter's center frequency, P1 selects notch bandwidth. Notch depth is fixed at about 60 dB.

4.5 MHz NOTCH FILTER

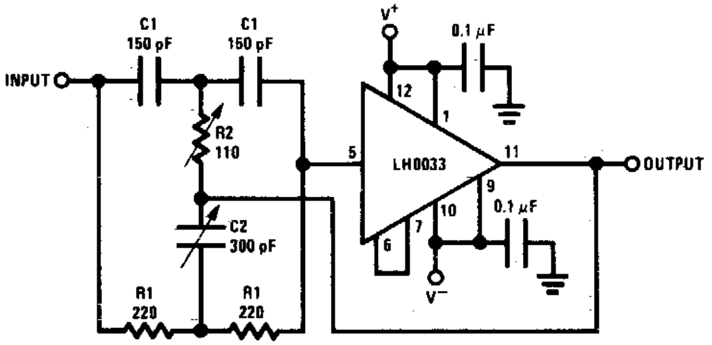
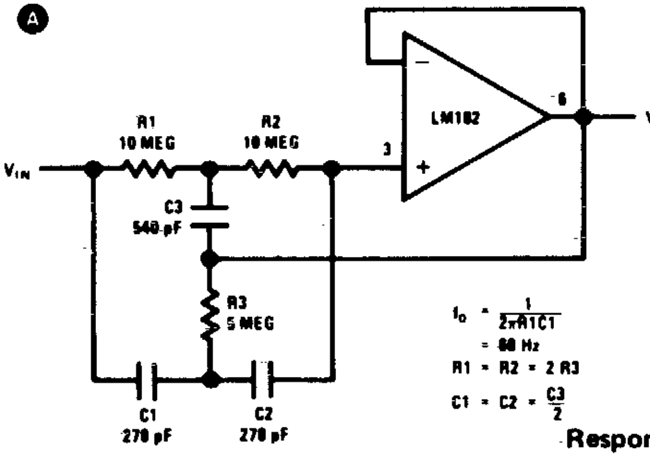

C1
150 pF
C1
150 pF
INPUT
R2
110
C2
300 pF
R1
220
R1
220
V+
0.1 μF
12
1
5
LH0033
11
9
10
7
6
0.1 μF
V−
OUTPUT

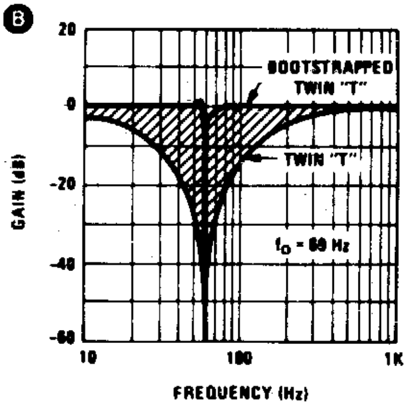
Fig. 33-8

Circuit Notes

Component value sensitivity is extremely critical, as are temperature coefficients and matching of the components. Best performance is attained when perfectly matched components are used and when the gain of the amplifier is unity. To illustrate, the quality factor Q is very high as amplifier gain approaches 1 with all components matched (in fact, theoretically it approaches ∞) but decreases to about 12.5 with the amplifier gain at 0.98.

HIGH Q NOTCH FILTER


A
R1
10 MEG
R2
10 MEG
V_IN
C3
540 pF
R3
5 MEG
C1
270 pF
C2
270 pF
LM102
3
6
V_OUT
f_o = 1/(2πR1C1)
= 60 Hz
R1 = R2 = 2 R3
C1 = C2 = C3/2



Response of High and Low Q Notch Filter

Fig. 33-9

Circuit Notes

A shows a twin-T network connected to an LM102 to form a high Q, 60 Hz notch filter. The junction of R3 and C3, which is normally connected to ground, is bootstrapped to the output of the follower. Because the output of the follower is a very low impedance, neither the depth nor the frequency of the notch change; however, the Q is raised in proportion to the amount of signal fed back to R3 and C3. B shows the response of a normal twin-T and the response with the follower added.

REJECTION FILTER

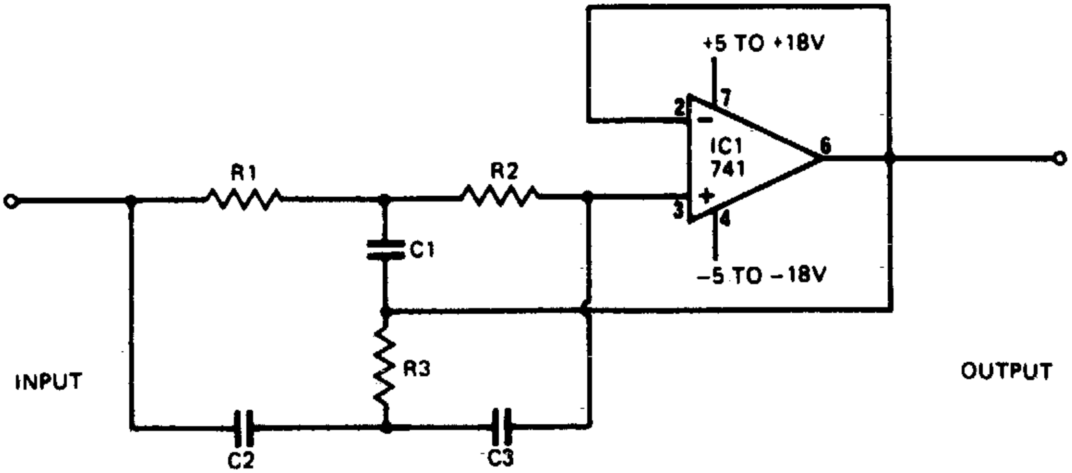

+5 TO +18V
7
2
IC1
741
6
R1
R2
3
4
C1
–5 TO –18V
INPUT
R3
OUTPUT
C2
C3

Fig. 33-10

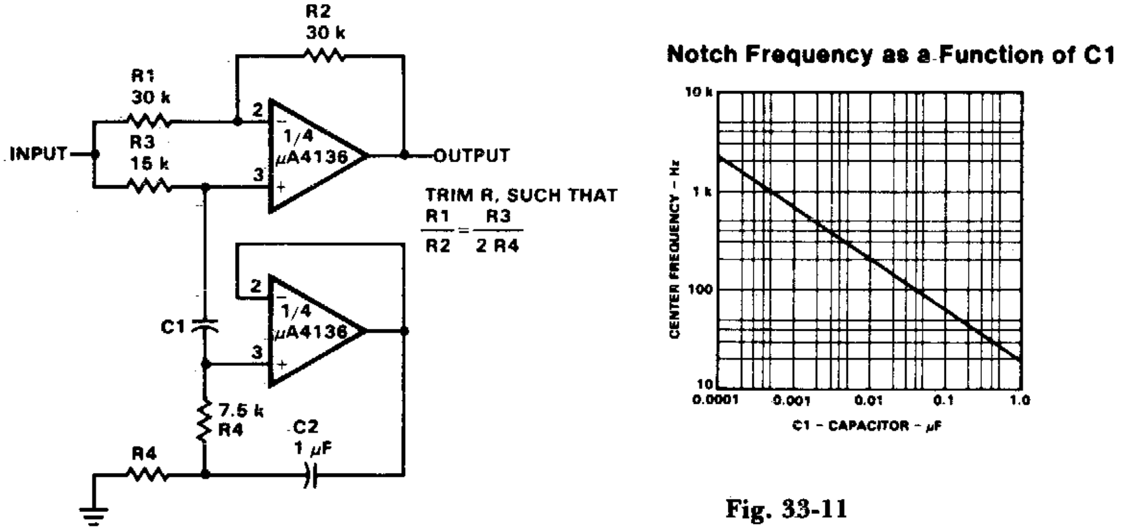
Circuit Notes

This narrowband filter using the 741 operational amplifier can provide up to 60 dB of rejection. With resistors equal to 100 K and capacitors equal to 320 pF, the circuit will reject 50 Hz. Frequencies within the range 1 Hz to 10 kHz may be rejected by selecting components in accordance with the formula:

$$F = \frac{1}{2\pi RC}$$

To obtain rejections better than 40 dB, resistors should be matched to 0.1% and capacitors to 1%.

NOTCH FILTER USING THE μA4136 AS A GYRATOR


R2
30 k
R1
30 k
R3
15 k
INPUT
2
3
1/4
μA4136
OUTPUT
2
3
1/4
μA4136
C1
7.5 k
R4
C2
1 μF
R4
10 k
1 k
100
10
CENTER FREQUENCY – Hz
0.0001
0.001
0.01
0.1
1.0
C1 – CAPACITOR – μF

TRIM R, SUCH THAT $\frac{R1}{R2} = \frac{R3}{2\,R4}$

Notch Frequency as a Function of C1

Fig. 33-11

1 kHz BANDPASS ACTIVE FILTER

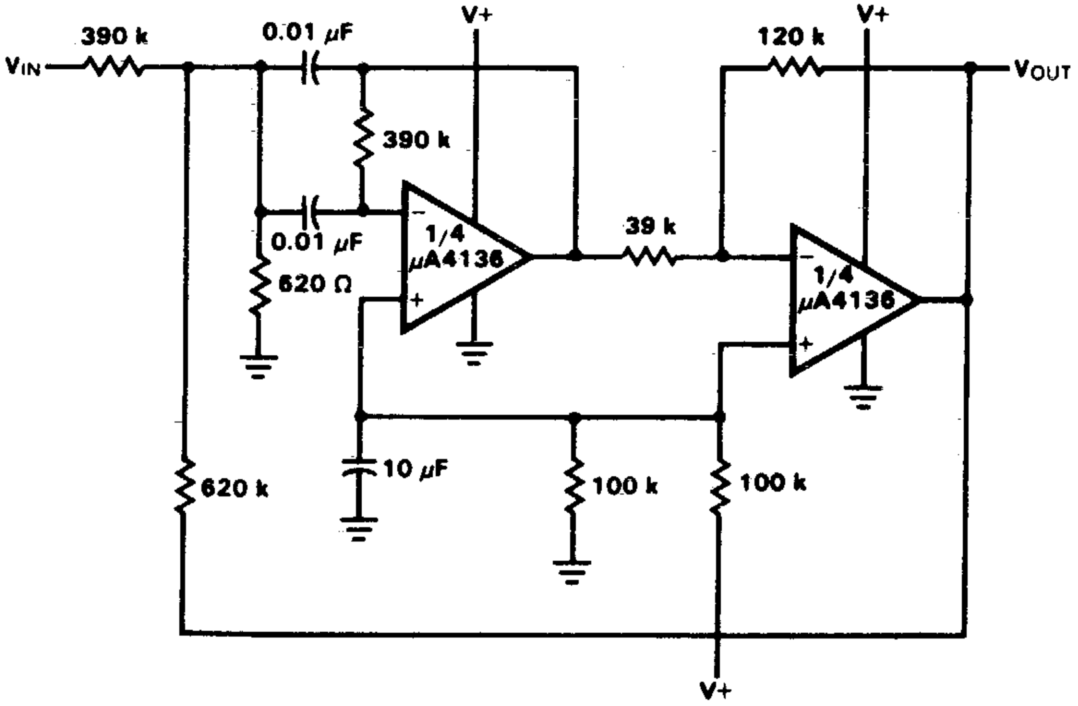
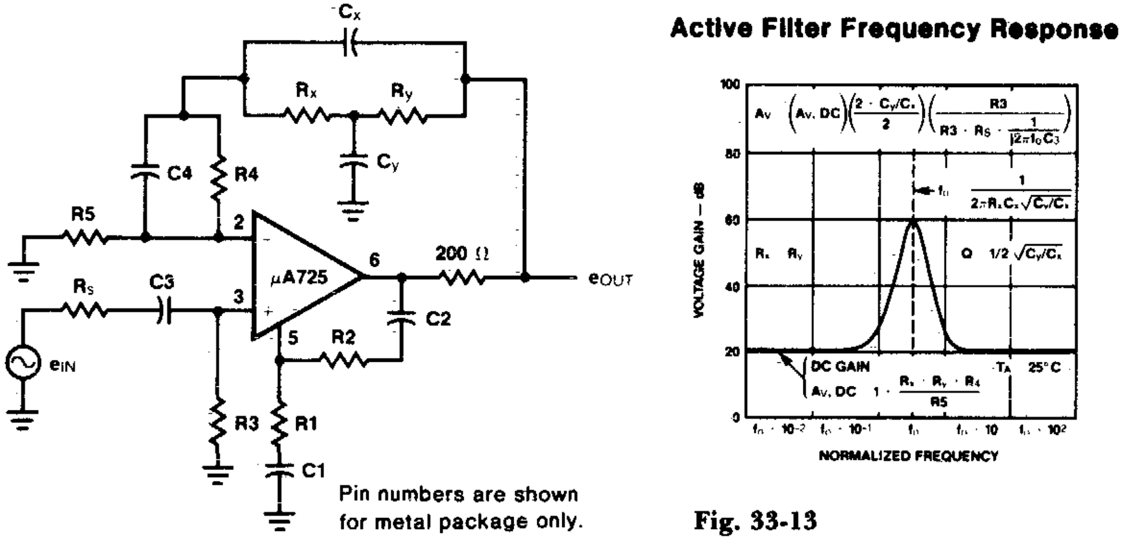

VIN
390 k
0.01 µF
V+
390 k
0.01 µF
620 Ω
1/4
µA4136
620 k
10 µF
100 k
39 k
120 k
V+
VOUT
1/4
µA4136
100 k
V+

Fig. 33-12

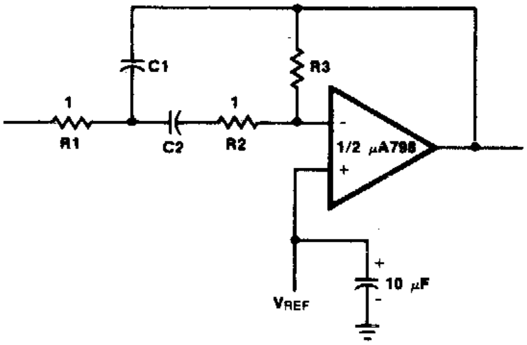
BANDPASS ACTIVE FILTER WITH 60 dB GAIN


Cx
Rx
Ry
Cy
C4
R4
R5
2
µA725
6
200 Ω
eOUT
RS
C3
3
5
R2
C2
eIN
R3
R1
C1
Active Filter Frequency Response
AV (AV, DC)((2 · Cy/Cx)/2)(R3/(R3 · RS · 1/(j2πf0C3)))
f0 1/(2πRxCx√(Cy/Cx))
Rx Ry
Q 1/2 √(Cy/Cx)
DC GAIN
AV, DC 1 · (Rx · Ry · R4)/R5
TA 25°C
VOLTAGE GAIN — dB
100
80
60
40
20
0
f0 · 10-2
f0 · 10-1
f0
f0 · 10
f0 · 102
NORMALIZED FREQUENCY

Pin numbers are shown for metal package only.

Fig. 33-13

MULTIPLE FEEDBACK BANDPASS FILTER



C1
R3
1
1
R1
C2
R2
1/2 μA798
-
+
+
10 μF
-
V_REF

$f_o \overset{\Delta}{=}$ center frequency

$BW \overset{\Delta}{=}$ Bandwidth

R in kΩ

C in μF

$Q = \frac{f_o}{BW} < 10$

$C1 = C2 = \frac{Q}{3}$

$\left.\begin{array}{l} R1 = R2 = 1 \\ R3 = 9Q^2 - 1 \end{array}\right\}$ Use scaling factors in these expressions

If source impedance is high or varies, filter may be preceded with voltage follower buffer to stabilize filter parameters.

Design example:

given: $Q = 5$, $f_o = 1$ kHz

Let $R1 = R2 = 10$ kΩ

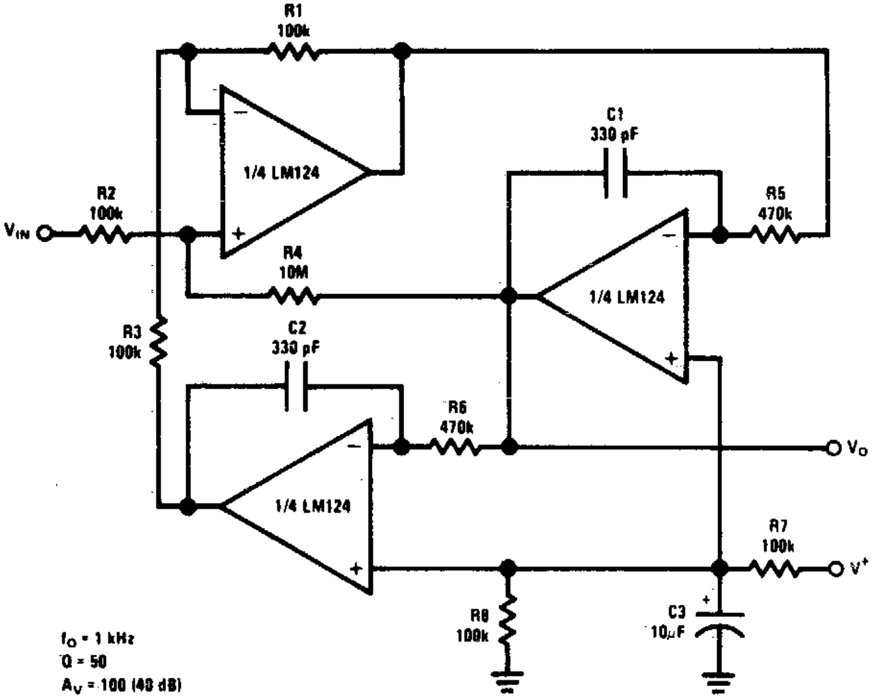
then $R3 = 9(5)^2 - 10$

$R3 = 215$ kΩ

$C = \frac{5}{3} = 1.6$ nF

Fig. 33-14

BIQUAD RC ACTIVE BANDPASS FILTER



R1
100k
C1
330 pF
1/4 LM124
R2
100k
V_IN
R5
470k
R4
10M
1/4 LM124
R3
100k
C2
330 pF
R6
470k
V_O
1/4 LM124
R7
100k
V+
R8
100k
C3
10μF

$f_O = 1$ kHz

$Q = 50$

$A_V = 100$ (40 dB)

Fig. 33-15

400 Hz LOW-PASS BUTTERWORTH ACTIVE FILTER

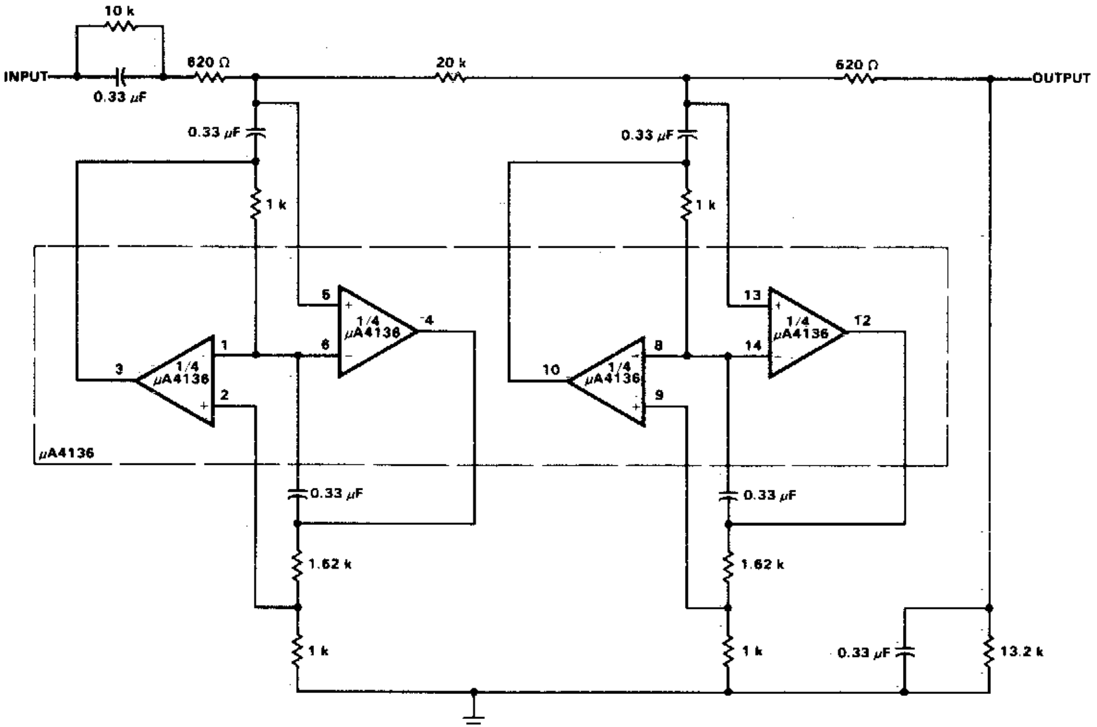

10 k
INPUT
0.33 μF
620 Ω
20 k
620 Ω
OUTPUT
0.33 μF
0.33 μF
1 k
1 k
5
6
4
1/4 μA4136
1
3
2
1/4 μA4136
13
14
12
1/4 μA4136
8
10
9
1/4 μA4136
μA4136
0.33 μF
0.33 μF
1.62 k
1.62 k
1 k
1 k
0.33 μF
13.2 k

Fig. 33-16

VARIABLE BANDWIDTH BANDPASS ACTIVE FILTER

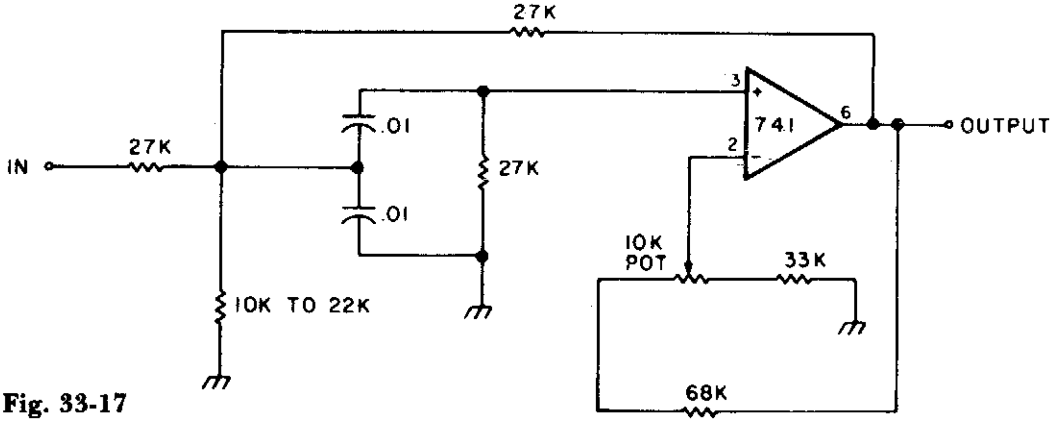

27K
IN
27K
.01
.01
27K
3
2
6
741
OUTPUT
10K POT
33K
10K TO 22K
68K

Fig. 33-17

Circuit Notes

This circuit has adjustable bandwidth with values for a center frequency of about 800 Hz. The 10 K pot adjusts bandwidth from approximately ±350 Hz to ±140 Hz at 3 dB down points.

LOW-PASS FILTER

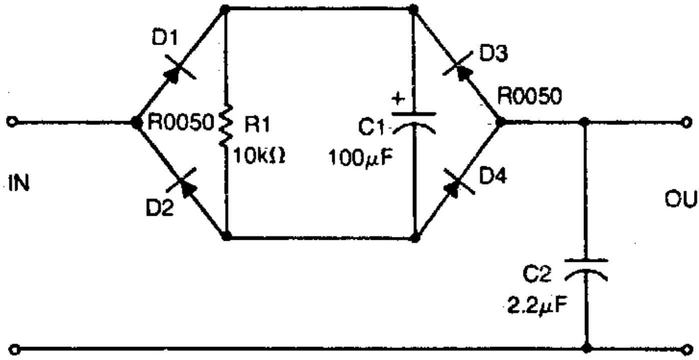

D1
D3
R0050
R0050
R1
10kΩ
C1
100μF
+
D2
D4
IN
OUT
C2
2.2μF

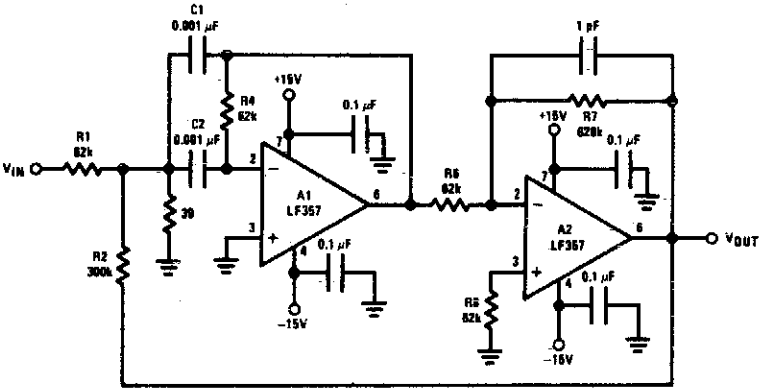
Fig. 33-18

D1, D2, D3, D4—HEP R0050
C1—100μF, 50V electrolytic
C2—2.2μF
R1—10kΩ, 1/2W

Circuit Notes

This nonlinear, passive filter circuit rejects ripple (or unwanted but fairly steady voltage) without appreciably affecting the rise time of a signal. The circuit works best when the signal level is considerably lower than the unwanted ripple, provided the ripple level is fairly constant. The circuit has characteristics similar to two peak-detecting sample-and-hold circuits in tandem with a voltage averager.

HIGH Q BANDPASS FILTER


C1
0.001 μF
C2
0.001 μF
R4
62k
R1
62k
V_IN
39
R2
300k
+15V
0.1 μF
7
2
3
6
4
A1
LF357
0.1 μF
−15V
R6
62k
1 pF
+15V
R7
620k
0.1 μF
A2
LF357
R5
62k
0.1 μF
−15V
V_OUT

- By adding positive feedback (R2) Q increases to 40
- $f_{BP} = 100$ kHz

$$\frac{V_{OUT}}{V_{IN}} = 10\sqrt{Q}$$

- Clean layout recommended
- Response to a 1 Vp-p tone burst: 300 μs

Fig. 33-19

MFB BANDPASS FILTER FOR MULTICHANNEL TONE DECODER

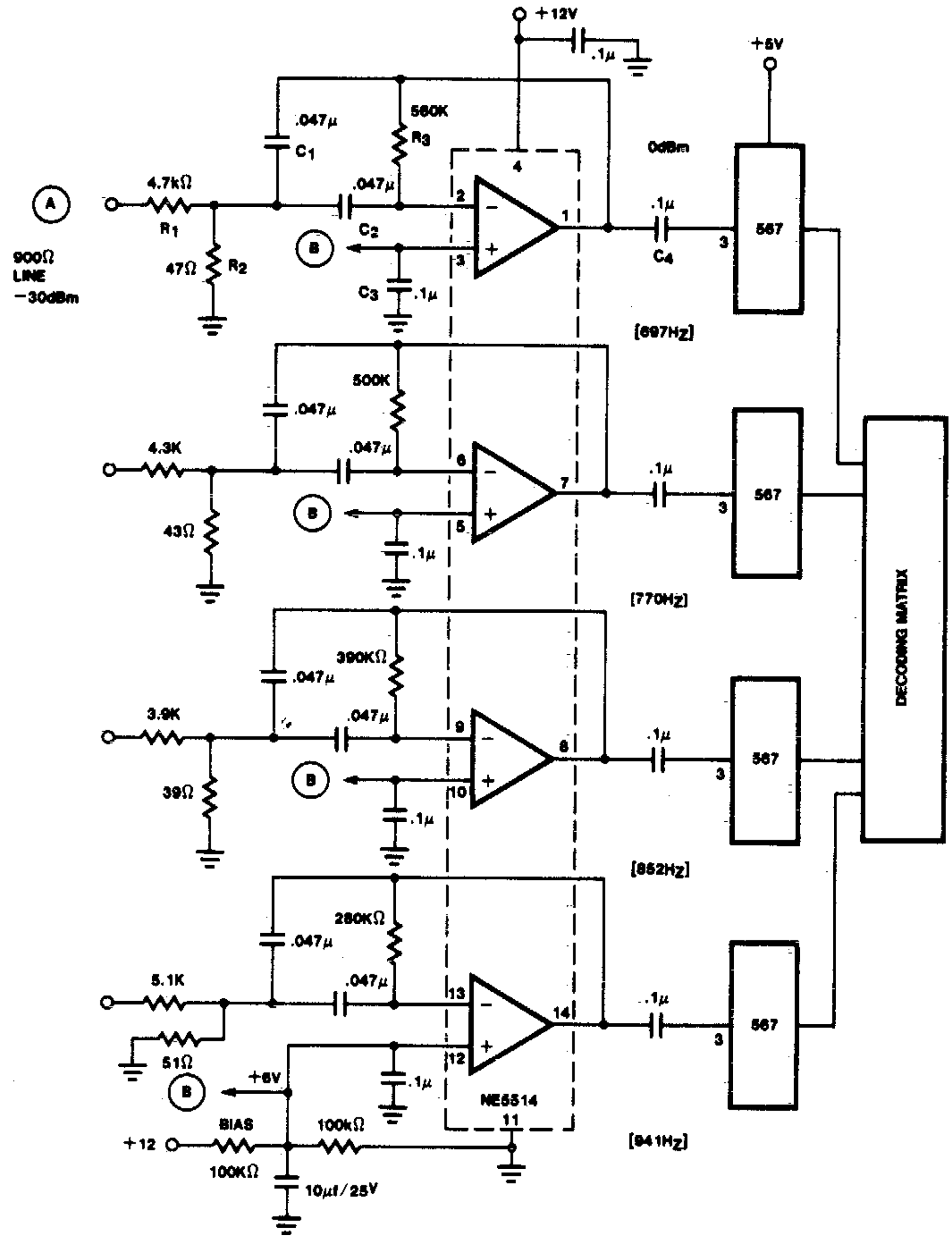
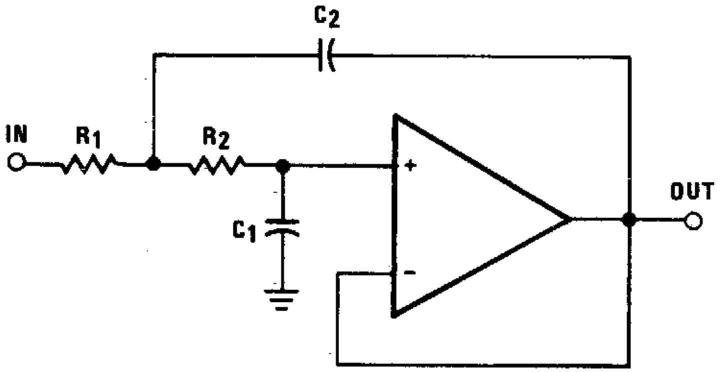

+12V
.1μ
+5V
.047μ
C1
560K
R3
4
0dBm
A
4.7kΩ
R1
.047μ
C2
2
1
.1μ
C4
3
567
B
3
900Ω
LINE
−30dBm
47Ω
R2
C3
.1μ
[697Hz]
500K
.047μ
4.3K
.047μ
6
7
.1μ
3
567
B
5
43Ω
.1μ
[770Hz]
DECODING MATRIX
390KΩ
.047μ
3.9K
.047μ
9
8
.1μ
3
567
B
10
39Ω
.1μ
[852Hz]
280KΩ
.047μ
5.1K
.047μ
13
14
.1μ
3
567
51Ω
12
.1μ
+6V
B
NE5514
11
BIAS
100kΩ
+12
100KΩ
10μf/25V
[941Hz]

Fig. 33-20

SALLEN-KEY SECOND ORDER LOW-PASS FILTER

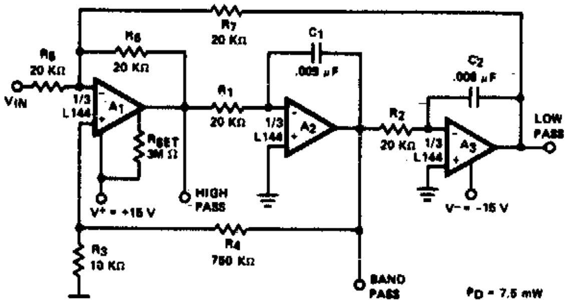

C2
IN
R1
R2
C1
+
-
OUT

NOTES:

1. Make $R_1 = R_2$
2. $fc = \frac{1}{2\ R_1\sqrt{C_1C_2}}$
3. $Q = \frac{1}{2}\sqrt{\frac{C_2}{C_1}}$

Fig. 33-21

THREE AMPLIFIER ACTIVE FILTER

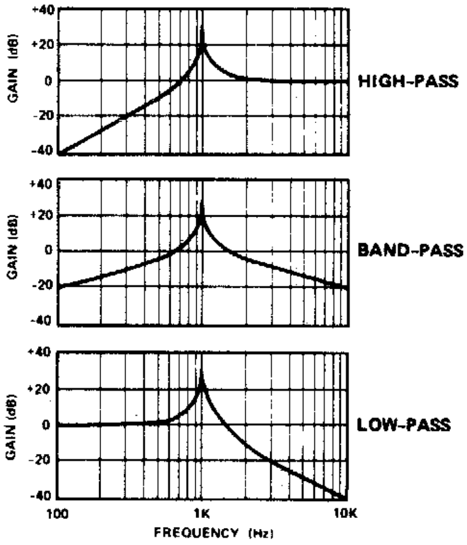

R7
20 KΩ
R6
20 KΩ
R5
20 KΩ
VIN
1/3
L144
A1
RSET
3M Ω
V+ = +15 V
HIGH
PASS
R1
20 KΩ
C1
.009 μF
1/3
L144
A2
R2
20 KΩ
C2
.008 μF
1/3
L144
A3
V− = −15 V
LOW
PASS
R3
10 KΩ
R4
750 KΩ
BAND
PASS
PD = 7.5 mW

$f_o = 1K\ Hz$
$Q = 26$
$H_o = 26$
$Q\ max < \frac{A_0}{5}$
$C_1 = C_2$
$R_5 = R_6 = R_7$

$\frac{R_4}{R_3} = 3\ H_o - 1\ f_o,\ H_o << \frac{A_0}{3}$

$R_2C_2 = \frac{H_o}{2\pi f_o Q}$

$R_1C_1 = \frac{Q}{2\pi f_o H_o}$



Bode plots of Active Filter Output

Fig. 33-22

Circuit Notes

The active filter is a state variable filter with bandpass, high-pass and low-pass outputs. It is a classical analog computer method of implementing a filter using three amplifiers and only two capacitors.

BANDPASS STATE VARIABLE FILTER

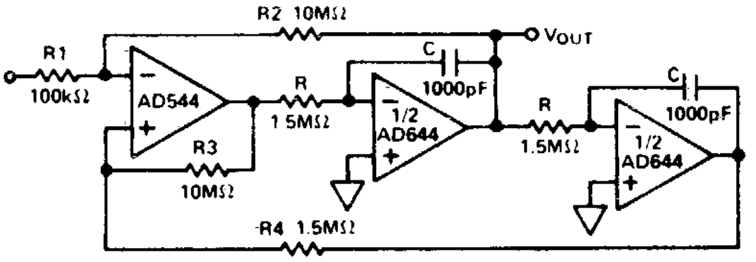

R2 10MΩ
VOUT
R1
100kΩ
AD544
R3
10MΩ
R
1.5MΩ
C
1000pF
1/2
AD644
R
1.5MΩ
C
1000pF
1/2
AD644
R4 1.5MΩ

Fig. 33-23

f_o = CENTER FREQUENCY = $1/2\pi RC$

Q_o = QUALITY FACTOR = $\frac{R_1 + R_2}{2R_1}$

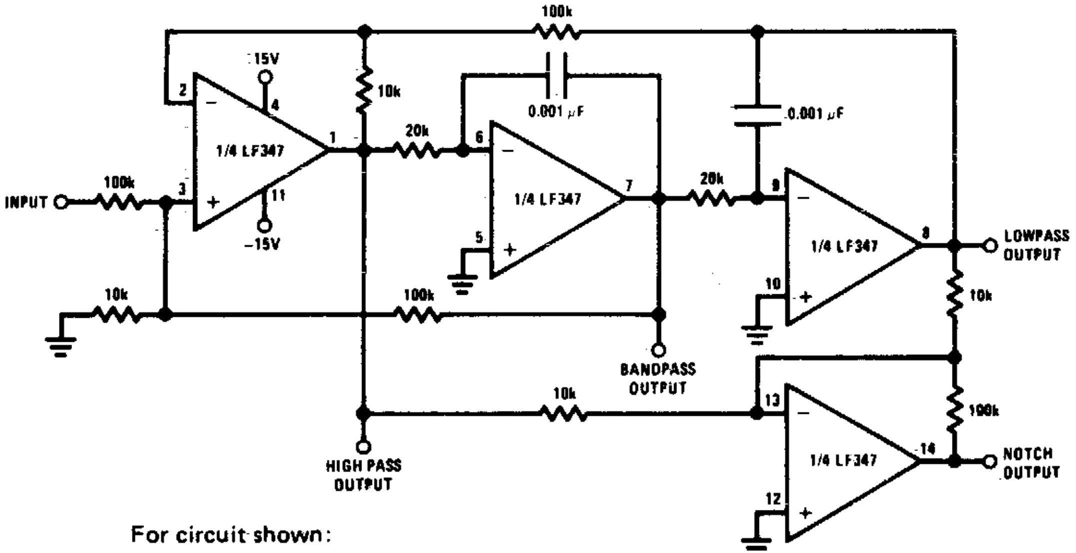
H_o = GAIN AT RESONANCE = R_2/R_1

$R_3 = R_4 \approx 10^8/f_o$

Qo, IS ADJUSTABLE BY VARYING R2

fo, IS ADJUSTABLE BY VARYING R OR C

UNIVERSAL STATE VARIABLE FILTER

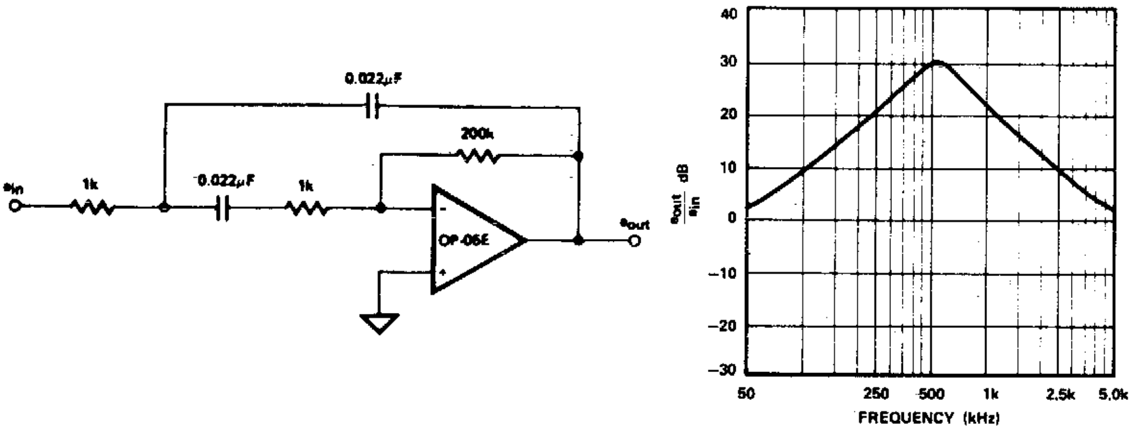

100k
+15V
10k
0.001 μF
0.001 μF
2
4
1/4 LF347
1
20k
6
INPUT
100k
3
11
−15V
1/4 LF347
7
20k
9
5
1/4 LF347
8
LOWPASS OUTPUT
10k
100k
10
10k
BANDPASS OUTPUT
10k
13
100k
1/4 LF347
14
NOTCH OUTPUT
12
HIGH PASS OUTPUT

For circuit shown:
f_O = 3 kHz, f_{NOTCH} = 9.5 kHz
Q = 3.4
Passband gain:
    Highpass – 0.1
    Bandpass – 1
    Lowpass – 1
    Notch – 10

- $f_O \times Q \leq 200$ kHz
- 10V peak sinusoidal output swing without slew limiting to 200 kHz
- See LM348 data sheet for design equations

Fig. 33-24

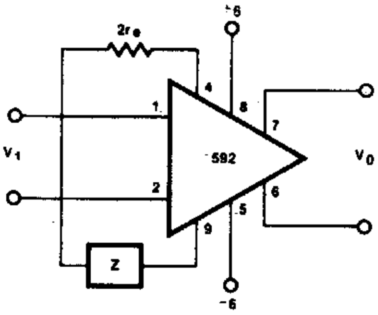
500 Hz SALLEN-KEY BANDPASS FILTER



0.022µF
200k
e_{in}
1k
0.022µF
1k
OP-06E
e_{out}
40
30
20
10
0
−10
−20
−30
$\frac{e_{out}}{e_{in}}$ dB
50
250
500
1k
2.5k
5.0k
FREQUENCY (kHz)

Fig. 33-25

FILTER NETWORKS

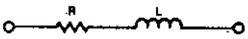
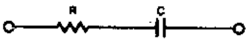
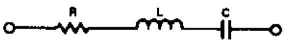
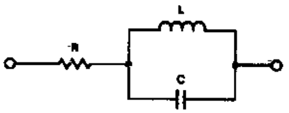


$2r_e$
+6
4
8
1
7
V_1
592
V_0
2
6
5
9
Z
−6

$$\frac{V_0(s)}{V_1(s)} \cong \frac{1.4 \times 10^4}{Z(s) + 2r_e}$$

$$\cong \frac{1.4 \times 10^4}{Z(s) + 32}$$

BASIC CONFIGURATION

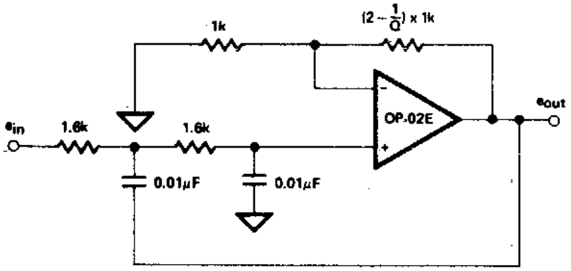
Z NETWORK	FILTER TYPE	$\frac{V_0(s)}{V_1(s)}$ TRANSFER FUNCTION
	LOW PASS	$\frac{1.4 \times 10^4}{L}\left[\frac{1}{s + R/L}\right]$
	HIGH PASS	$\frac{1.4 \times 10^4}{R}\left[\frac{s}{s + 1/RC}\right]$
	BAND PASS	$\frac{1.4 \times 10^4}{L}\left[\frac{s}{s^2 + R/L\, s + 1/LC}\right]$
	BAND REJECT	$\frac{1.4 \times 10^4}{R}\left[\frac{s^2 + 1/LC}{s^2 + 1/LC + s/RC}\right]$

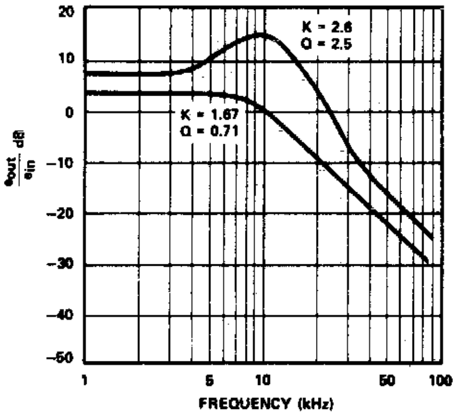
NOTE

In the networks above, the R value used is assumed to include $2r_e$, or approximately 32Ω.

Fig. 33-26

EQUAL COMPONENT SALLEN-KEY LOW-PASS FILTER

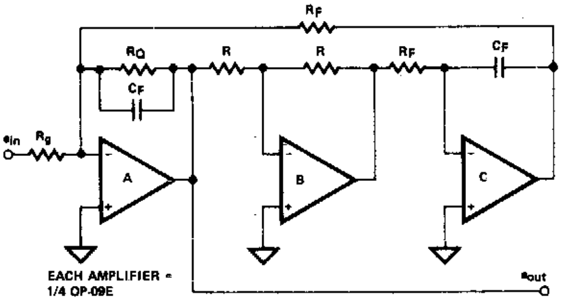

1k
$(2 - \frac{1}{Q}) \times 1k$
e_{in}
1.6k
1.6k
0.01μF
0.01μF
OP-02E
e_{out}



Equal R, Equal C Sallen-Key Response

Fig. 33-27

BIQUAD FILTER

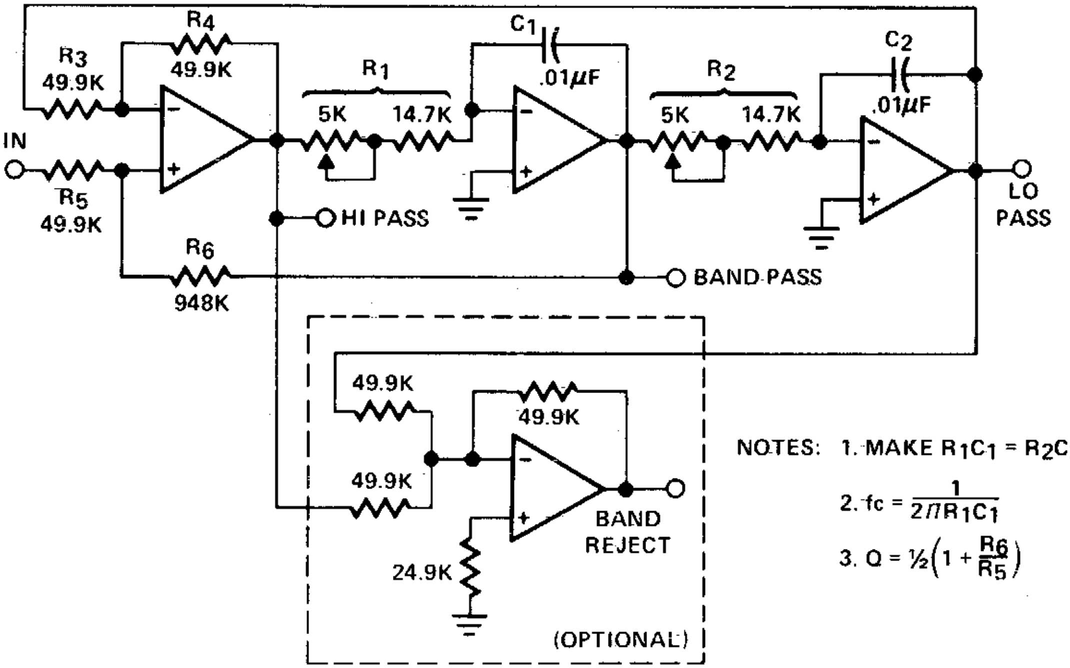

R_F
R_Q
R
R
R_F
C_F
C_F
e_{in}
R_g
A
B
C
EACH AMPLIFIER = 1/4 OP-09E
e_{out}

Circuit Notes

The biquad filter, while appearing very similar to the state-variable filter, has a bandwidth that is fixed regardless of center frequency. This type of filter is useful in applications such as spectrum analyzers, which require a filter with a fixed bandwidth.

Fig. 33-28

SECOND ORDER STATE VARIABLE FILTER (1 kHz, Q = 10)


R4
49.9K
R3
49.9K
C1
.01μF
R1
5K
14.7K
R2
5K
14.7K
C2
.01μF
IN
R5
49.9K
HI PASS
LO
PASS
R6
948K
BAND PASS
49.9K
49.9K
49.9K
24.9K
BAND
REJECT
(OPTIONAL)

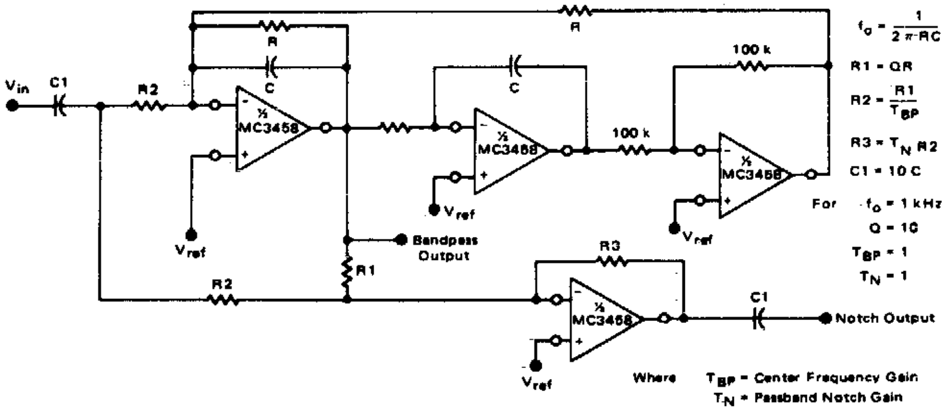
NOTES: 1. MAKE $R_1C_1 = R_2C_2$

2. $f_c = \frac{1}{2\pi R_1C_1}$

3. $Q = \frac{1}{2}\left(1 + \frac{R_6}{R_5}\right)$

Fig. 33-29

BIQUAD FILTER


R
R
C
C
100 k
V_{in}
C1
R2
½ MC3458
½ MC3458
100 k
½ MC3458
V_{ref}
V_{ref}
V_{ref}
Bandpass Output
R1
R2
R3
½ MC3458
C1
Notch Output
V_{ref}

$f_o = \frac{1}{2\pi RC}$

$R1 = QR$

$R2 = \frac{R1}{T_{BP}}$

$R3 = T_N R2$

$C1 = 10\,C$

$V_{ref} = \frac{1}{2} V_{CC}$

For $f_o = 1$ kHz

$Q = 10$

$T_{BP} = 1$

$T_N = 1$

Where T_{BP} = Center Frequency Gain

T_N = Passband Notch Gain

$R = 160\ k\Omega$

$C = 0.001\ \mu F$

$R1 = 1.6\ M\Omega$

$R2 = 1.6\ M\Omega$

$R3 = 1.6\ M\Omega$

Fig. 33-30

TUNABLE ACTIVE FILTER

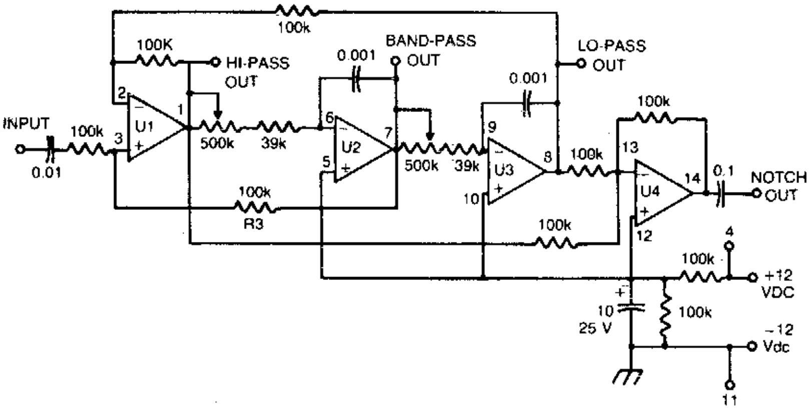
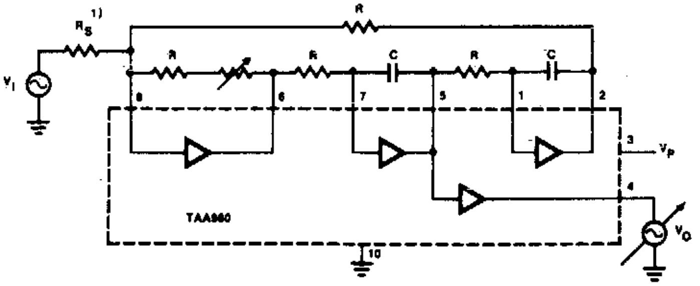

100k
100K
HI-PASS OUT
0.001
BAND-PASS OUT
LO-PASS OUT
0.001
100k
INPUT
100k
0.01
U1
500k
39k
U2
500k
39k
U3
100k
U4
NOTCH OUT
0.1
100k
R3
100k
100k
+12 VDC
10
25 V
100k
−12 Vdc
4
11
1
2
3
5
6
7
8
9
10
12
13
14

Fig. 33-31

Circuit Notes

The high-pass and low-pass outputs covering the range of 300 Hz to 3000 Hz have been summed in the fourth op amp to provide a notch output. The potentiometers must have a reverse log taper. Fixed-frequency active filter center frequency is 1 kHz, with a Q of 50.

ACTIVE RC FILTER FOR FREQUENCIES UP TO 150 kHz


R_S 1)
R
V_i
R
R
C
R
C
8
6
7
5
1
2
3
V_P
4
TAA960
10
V_O

R = 10kΩ

This frequency range can be extended to 200kHz if a feed forward capacitor is connected between pin 5 and 8.

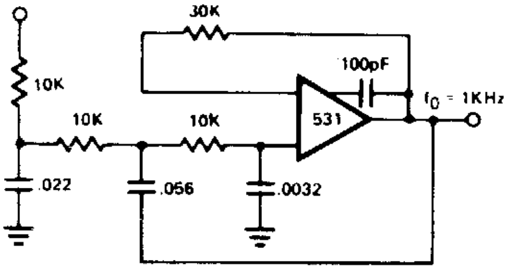
Fig. 33-32

f	Frequency	$\frac{1}{2\pi RC}$	
V_P	Supply voltage	6	V
	Filter performance		
Q	at $T_A = 25°C$	40 to 55	
Q	at $T_A = -30$ to $+65°C$	35 to 55	
V_i	Input voltage	400	mV
V_o	Output voltage	400	mV
d_{tot}	Distortion at $V_o = 350mV$	2	%
S/N	S/N ratio at $V_o = 400mV$	50	dB
R_S	Input resistor*	470	kΩ

*NOTE

Value of input resistor to be determined for $\frac{V_o}{V_i} = 0.90$ to 1.1.

POLE ACTIVE LOW-PASS FILTER
(BUTTERWORTH MAXIMALLY FLAT RESPONSE)



30K
100pF
10K
10K
10K
531
f_0 = 1KHz
.022
.056
.0032

*Reference—EDN Dec. 15, 1970
Simplify 3-Pole Active Filter Design
A. Paul Brokow

RESPONSE OF 3-POLE ACTIVE BUTTERWORTH MAXIMALLY FLAT FILTER

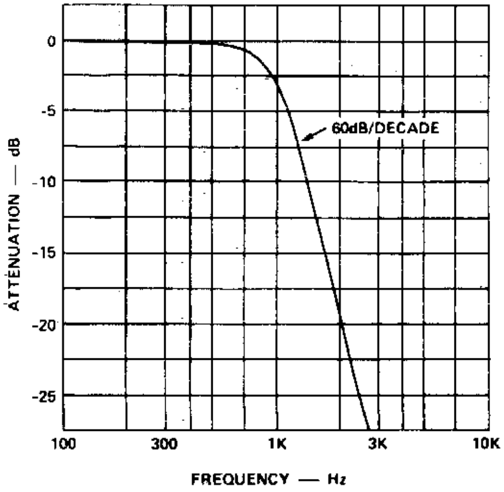
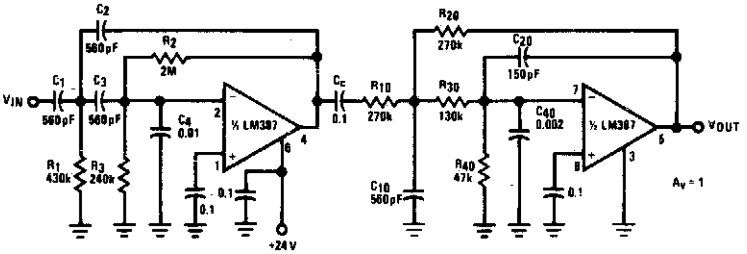
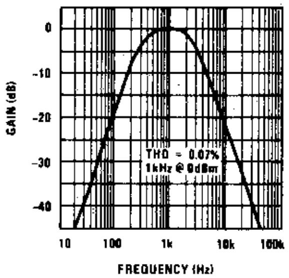


Fig. 33-33

SPEECH FILTER (300 Hz .3 kHz BANDPASS)



C2
560pF
R2
2M
V_{IN}
C1
560pF
C3
560pF
C4
0.01
R1
430k
R3
240k
2
½ LM307
1
6
4
0.1
0.1
+24V
C_c
0.1
R10
270k
R20
270k
R30
130k
C20
150pF
C10
560pF
R40
47k
C40
0.002
7
½ LM307
8
5
3
0.1
V_{OUT}
A_v = 1



Speech Filter Frequency Response

Fig. 33-34

0.1 Hz TO 10 Hz BANDPASS FILTER

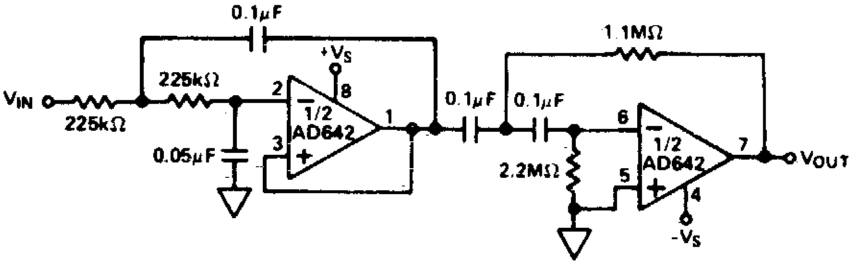
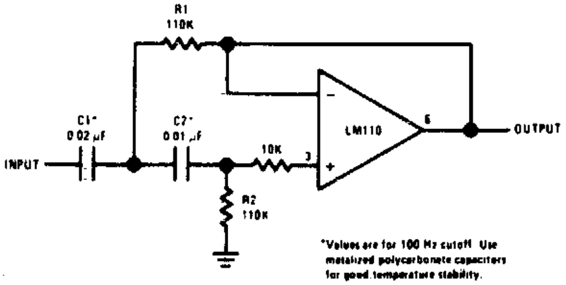

0.1μF
+VS
8
1.1MΩ
VIN
225kΩ
225kΩ
2
1/2
AD642
1
0.1μF
0.1μF
6
3
0.05μF
1/2
AD642
7
VOUT
2.2MΩ
5
4
-VS

Fig. 33-35

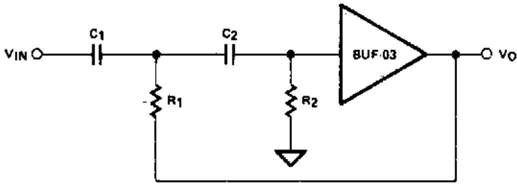
HIGH-PASS ACTIVE FILTER


R1
110K
C1*
0.02 μF
C2*
0.01 μF
INPUT
10K
3
LM110
6
OUTPUT
R2
110K

*Values are for 100 Hz cutoff. Use metalized polycarbonate capacitors for good temperature stability.

Fig. 33-36

HIGH-PASS FILTER (HIGH FREQUENCY)


C1
C2
VIN
BUF-03
VOUT
R1
R2

$$\omega_o = \left(\frac{1}{R_1R_2C_1C_2}\right)^{½}$$

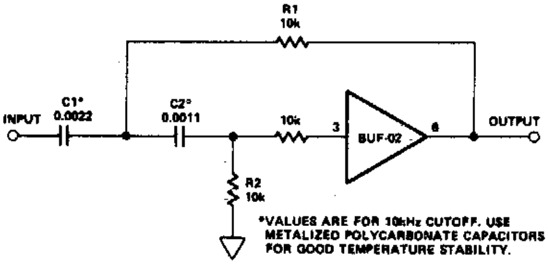
IF $C_1 = C_2 = C$, THEN

$$Q = \frac{(R_1/R_2)^{½}}{2}$$

f_o Hz	C pF	R_1 Ω	R_2 Ω	Q
500K	220	2.05K	1.02K	0.71

Fig. 33-38

SECOND ORDER HIGH-PASS ACTIVE FILTER


R1
10k
C1*
0.0022
C2*
0.0011
INPUT
10k
3
BUF-02
6
OUTPUT
R2
10k

*VALUES ARE FOR 10kHz CUTOFF. USE METALIZED POLYCARBONATE CAPACITORS FOR GOOD TEMPERATURE STABILITY.

Fig. 33-37

160 Hz BANDPASS FILTER

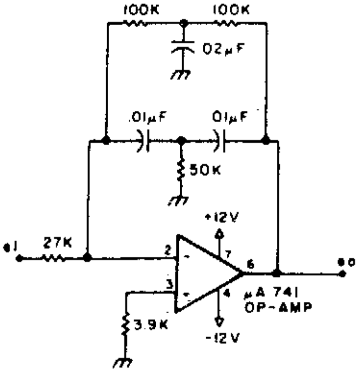
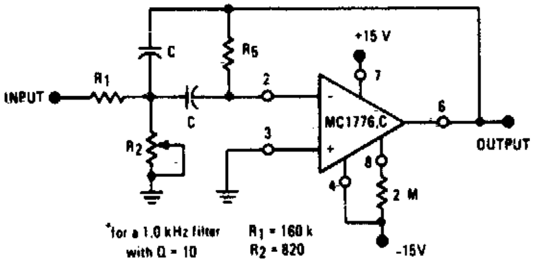

100K
100K
.02μF
.01μF
.01μF
50K
+12V
ei
27K
2
7
6
eo
3
4
μA 741
OP-AMP
3.9K
-12V

Fig. 33-39

MULTIPLE FEEDBACK BANDPASS FILTER (1.0 kHz)

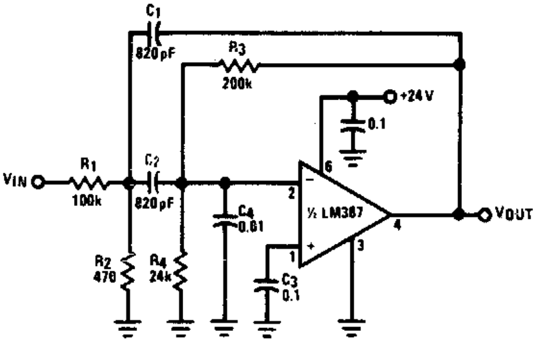

C
R5
+15 V
7
R1
INPUT
C
2
MC1776,C
6
OUTPUT
3
R2
8
4
2 M
-15V

*for a 1.0 kHz filter
with Q = 10
and A (f_0) = 1

R_1 = 160 k
R_2 = 820
R_5 = 300 k
C = 0.01 μF

Fig. 33-40

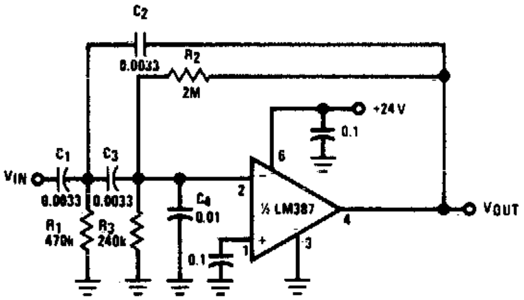
20 kHz BANDPASS ACTIVE FILTER


C1
820pF
R3
200k
+24V
0.1
VIN
R1
100k
C2
820pF
6
2
C4
0.01
½ LM387
4
VOUT
R2
470
R4
24k
1
C3
0.1
3

A_0 = −1
f_0 = 20kHz
Q = 10
THD < 0.1%

Fig. 33-41

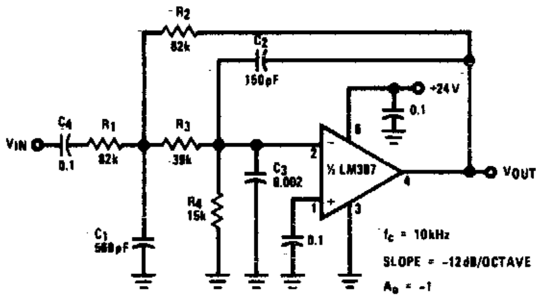
RUMBLE FILTER USING LM387


C2
0.0033
R2
2M
+24V
0.1
6
VIN
C1
C3
0.0033
0.0033
R1
470k
R3
240k
C4
0.01
2
½ LM387
4
VOUT
0.1
1
3

f_c = 50Hz
SLOPE = −12dB/OCTAVE
A_0 = −1
THD < 0.1%

Fig. 33-42

SCRATCH FILTER USING LM387


R2
82k
C2
150pF
+24V
0.1
VIN
C4
0.1
R1
82k
R3
39k
6
2
½ LM387
4
VOUT
C3
0.002
R4
15k
1
3
C1
560pF
0.1

f_c = 10kHz
SLOPE = −12dB/OCTAVE
A_0 = −1
THD < 0.1%

Fig. 33-43

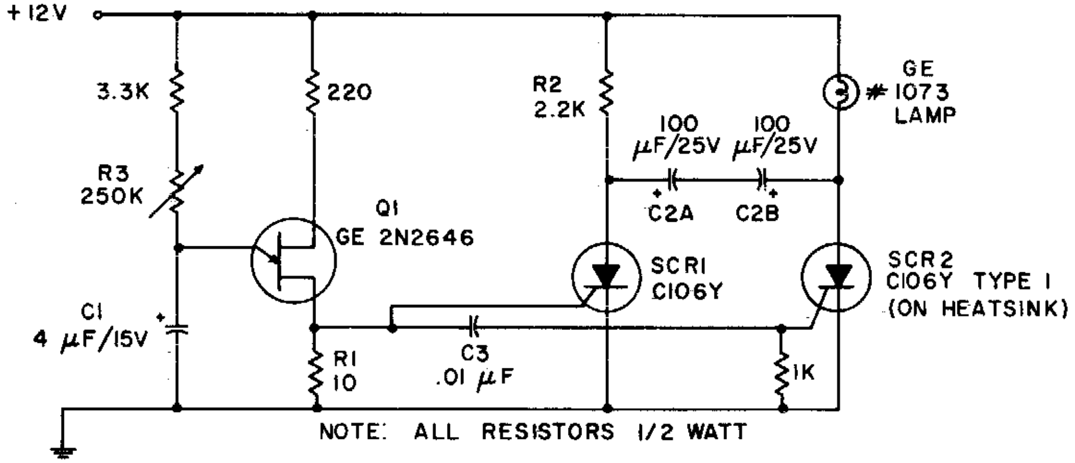
34

Flashers and Blinkers

The sources of the following circuits are contained in the Sources section beginning on page 730. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

Auto, Boat, or Barricade Flasher
Flip-Flop Flasher
Flashlight Finder
Low Frequency Lamp Flasher/Relay Driver
Low Cost Ring Counter
Ring Counter for Incandescent Lamps
Dual LED CMOS Flasher
Automatic Safety Flasher
Neon Blinker
Transistorized Flasher
Flasher/Light Control
Neon Tube Flasher
Dc Flasher with Adjustable On and Off Time
Low Voltage Flasher
1 A Lamp Flasher
Fast Blinker
3 V Flasher
Incandescent Bulb Flasher
Flasher for 4 Parallel LEDs
LED Booster
Safe, High Voltage Flasher
Alternating Flasher
Variable Flasher
Emergency Lantern/Flasher
High Efficiency Parallel Circuit Flasher
Minimum Power Flasher

AUTO, BOAT, OR BARRICADE FLASHER


+12V
3.3K
220
R2
2.2K
GE
# 1073
LAMP
100
µF/25V
100
µF/25V
C2A
C2B
R3
250K
Q1
GE 2N2646
SCR1
C106Y
SCR2
C106Y TYPE 1
(ON HEATSINK)
C1
4 µF/15V
R1
10
C3
.01 µF
1K

NOTE: ALL RESISTORS 1/2 WATT

Fig. 34-1

Circuit Notes

Because of its ability to withstand the heavy inrush currents, this incandescent lamp flasher uses the C106 SCR. With the components shown, the flash rate is adjustable by potentiometer R3 within the range of 36 flashes per minute to 160 flashes per minute.

FLIP-FLOP FLASHER

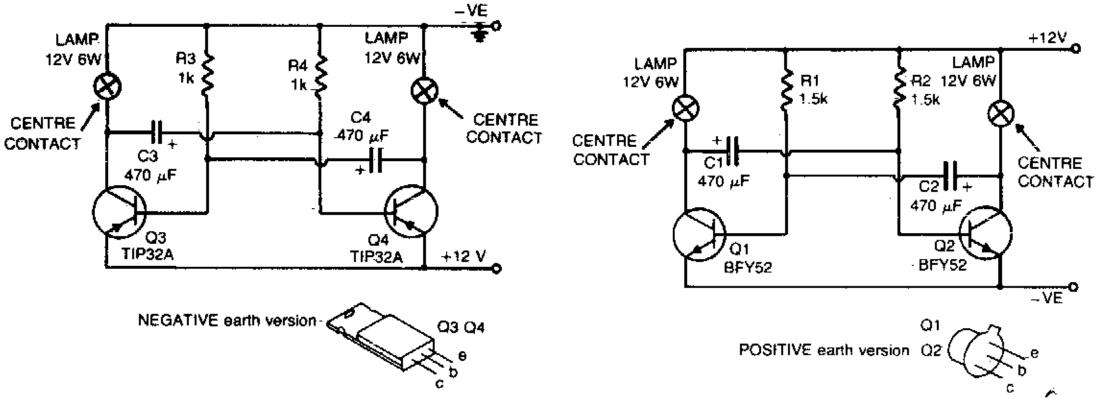
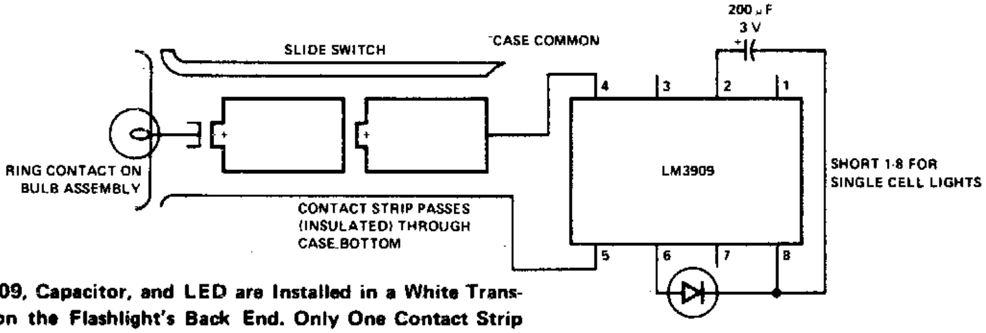

−VE
LAMP
12V 6W
R3
1k
R4
1k
LAMP
12V 6W
CENTRE
CONTACT
C4
470 µF
CENTRE
CONTACT
C3
470 µF
Q3
TIP32A
Q4
TIP32A
+12 V
NEGATIVE earth version
Q3 Q4
e
b
c
+12V
LAMP
12V 6W
R1
1.5k
R2
1.5k
LAMP
12V 6W
CENTRE
CONTACT
C1
470 µF
C2
470 µF
CENTRE
CONTACT
Q1
BFY52
Q2
BFY52
−VE
Q1
POSITIVE earth version
Q2
e
b
c

Fig. 34-2

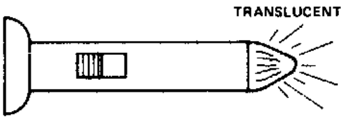
Circuit Notes

The flashing action is provided by a simple astable multivibrator timed to give a flashing rate of about 60 flashes for each lamp per minute. Circuit for positive earth systems uses NPN transistors. The other uses PNP transistors.

FLASHLIGHT FINDER


200 μF
3 V
SLIDE SWITCH
CASE COMMON
RING CONTACT ON BULB ASSEMBLY
CONTACT STRIP PASSES (INSULATED) THROUGH CASE BOTTOM
LM3909
SHORT 1-8 FOR SINGLE CELL LIGHTS
4
3
2
1
5
6
7
8

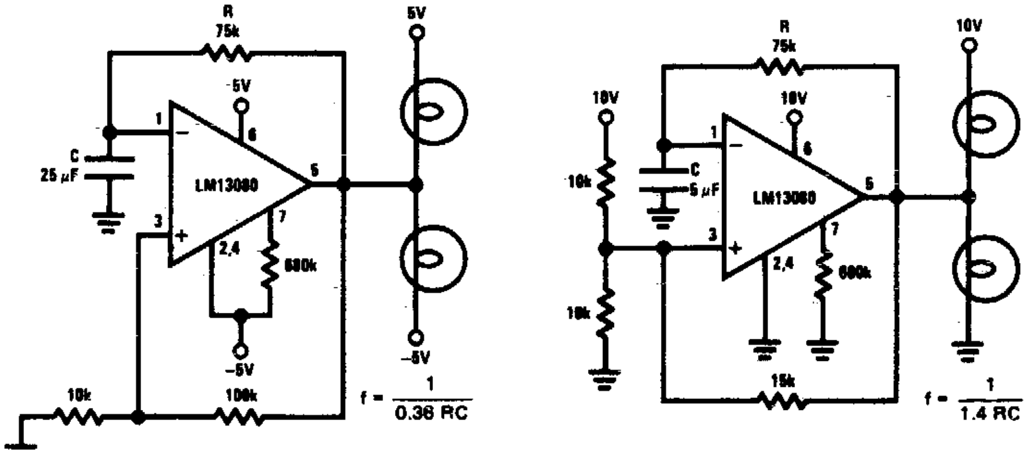
Note: LM3909, Capacitor, and LED are Installed in a White Translucent Cap on the Flashlight's Back End. Only One Contact Strip (in Addition to the Case Connection) is Needed for Flasher Power. Drawing Current Through the Bulb Simplifies Wiring and Causes Negligible Loss Since Bulb Resistance Cold is Typically Less than 2 Ω.


TRANSLUCENT

Note: Winking LED Inside, Locates Light in Total Darkness

Fig. 34-3

LOW FREQUENCY LAMP FLASHER/RELAY DRIVER


R
75k
5V
-5V
C
25 μF
1
6
3
5
7
2,4
LM13080
680k
-5V
-5V
10k
100k
f = 1/0.36 RC
R
75k
10V
10V
10V
10k
C
5 μF
1
6
3
5
7
2,4
LM13080
680k
10k
15k
f = 1/1.4 RC

a) Split Supplies

b) Single Supply

Fig. 34-4

Circuit Notes

This circuit is a low frequency warning device. The output of the oscillator is a square wave that is used to drive lamps or small relays. The circuit alternately flashes two incandescent lamps.

LOW COST RING COUNTER

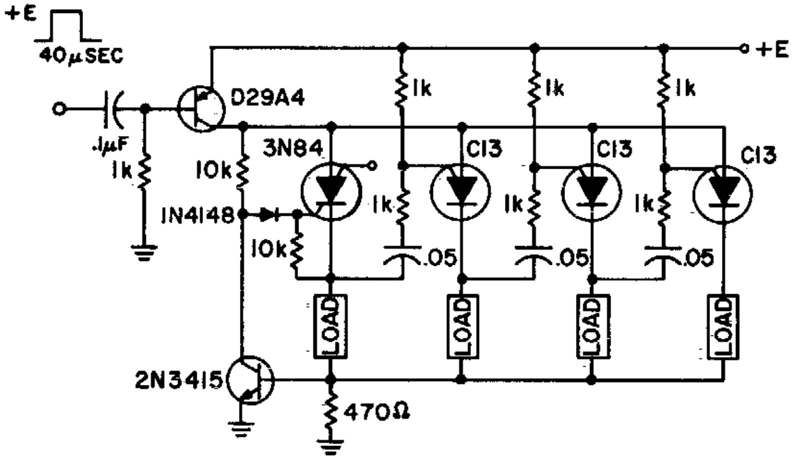

+E
40μSEC
.1μF
1k
D29A4
10k
3N84
1N4148
10k
1k
C13
1k
C13
1k
C13
1k
.05
.05
.05
+E
LOAD
LOAD
LOAD
LOAD
2N3415
470Ω

Fig. 34-5

Circuit Notes

This ring counter makes an efficient, low cost circuit featuring automatic resetting via the first stage 3N84. As many stages as desired may be cascaded.

RING COUNTER FOR INCANDESCENT LAMPS

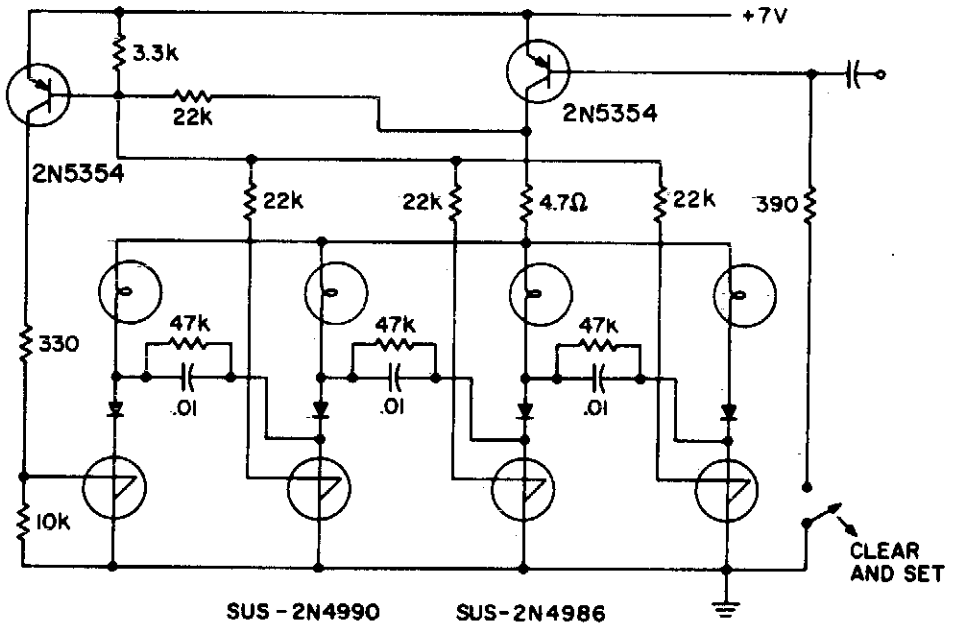
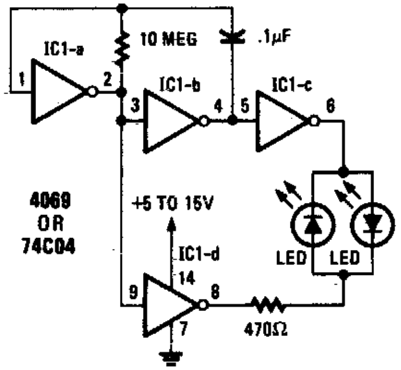

+7V
3.3k
22k
2N5354
2N5354
22k
22k
4.7Ω
22k
390
330
47k
47k
47k
.01
.01
.01
10k
CLEAR AND SET
SUS-2N4990
SUS-2N4986

Fig. 34-6

DUAL LED CMOS FLASHER



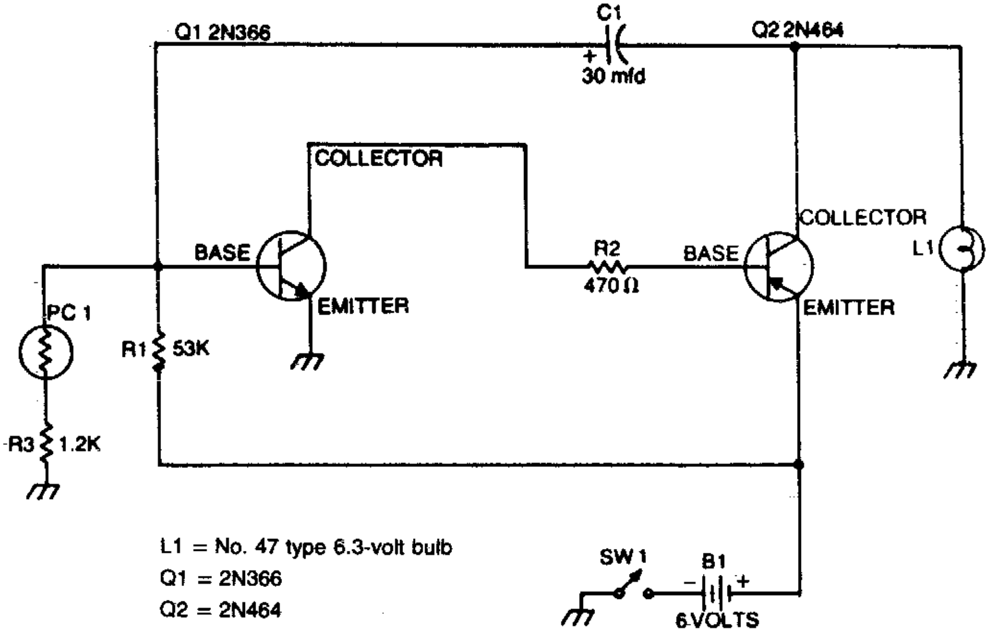
IC1-a
10 MEG
.1µF
IC1-b
IC1-c
1
2
3
4
5
6
4069
OR
74C04
+5 TO 15V
IC1-d
14
9
8
7
LED
LED
470Ω

Fig. 34-7

Circuit Notes

Inverters IC1-a and IC1-b form a multivibrator and IC1-c is a buffer. Inverter IC1-d is connected so that its output is opposite that of IC1-c; when pin 6 is high, then pin 8 is low and vice versa. Because pins 6 and 8 are constantly changing state, first one LED and then the other is on since they are connected in reverse. The light seems to jump back and forth between the LED's. The 470-ohm resistor limits LED current. Depending upon the supply voltage used, the value of the resistor may have to be changed to obtain maximum light output. To change the switching rate, change the value of the capacitor.

AUTOMATIC SAFETY FLASHER



Q1 2N366
C1
+
30 mfd
Q2 2N464
COLLECTOR
BASE
EMITTER
R2
470 Ω
BASE
COLLECTOR
EMITTER
L1
PC 1
R1
53K
R3
1.2K
SW 1
B1
6 VOLTS

L1 = No. 47 type 6.3-volt bulb
Q1 = 2N366
Q2 = 2N464

Fig. 34-8

Circuit Notes

This flasher only comes on at night. It furnishes a bright nighttime illumination, and shuts itself off automatically as soon as the sun comes up. The photocell must be mounted on top of the unit in such a way as to detect the greatest amount of available light.

NEON BLINKER

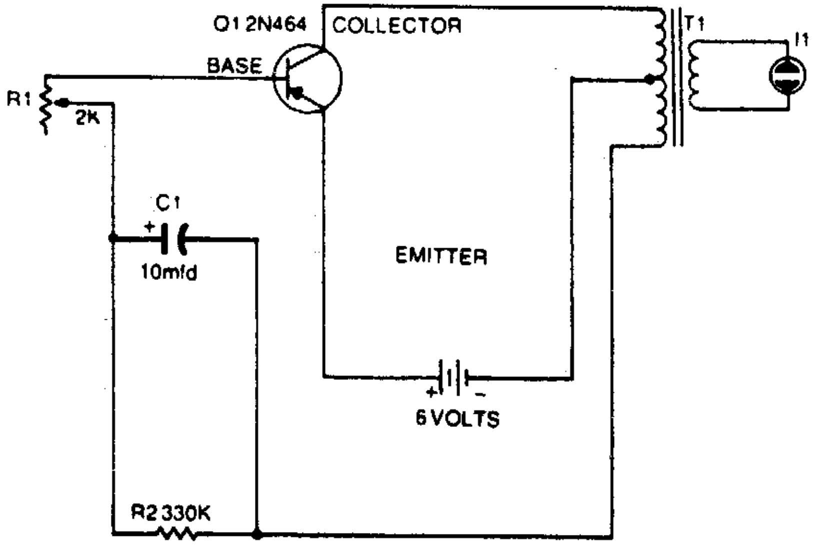
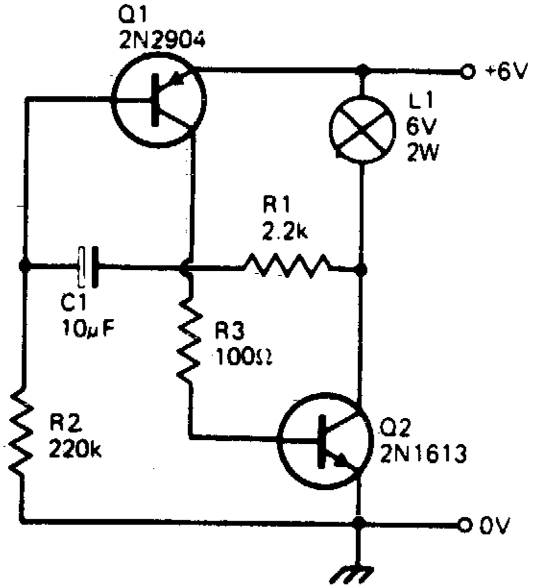

Q1 2N464
COLLECTOR
BASE
R1
2K
T1
I1
C1
+
10mfd
EMITTER
+
-
6 VOLTS
R2 330K

Fig. 34-9

Circuit Notes

The universal output transformer and the transistor form a low-frequency oscillator. The rate of flashing of the neon bulb is determined by potentiometer R1.

TRANSISTORIZED FLASHER

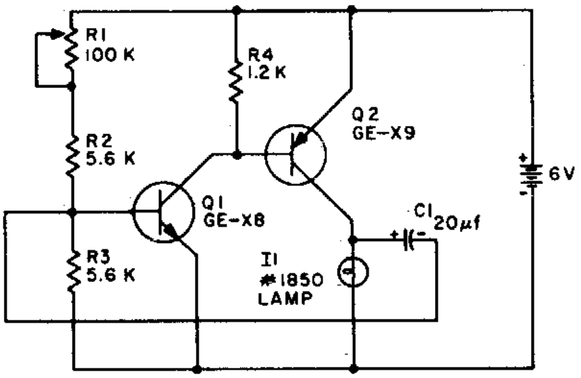

Q1
2N2904
+6V
L1
6V
2W
R1
2.2k
C1
10μF
R3
100Ω
R2
220k
Q2
2N1613
0V

Circuit Notes

This simple circuit will flash a 6 volt lamp at a rate determined by the size of capacitor C1. It is most economical on power as it only draws current when the lamp is on. When the lamp is off, both transistors are biased off.

Fig. 34-10

FLASHER/LIGHT CONTROL



R1
100 K
R4
1.2 K
Q2
GE-X9
R2
5.6 K
Q1
GE-X8
+
6V
C1
20µf
R3
5.6 K
I1
#1850
LAMP

Fig. 34-11

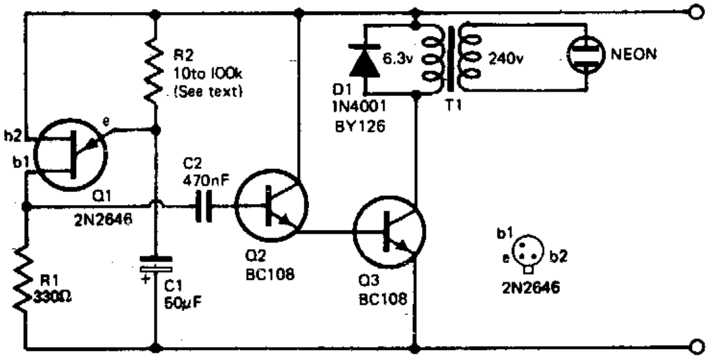
Parts List

C1 — 20-mfd, 6-volt electrolytic capacitor
I1 — 6-volt, GE No. 1850 lamp and socket
Q1 — GE-X8 transistor
Q2 — GE-X9 transistor
R1 — 100K-ohm, 2-watt potentiometer
R2, R3 — 5.6K-ohm, 1/2-watt resistor
R4 — 1.2K-ohm, 1/2-watt resistor
Battery — 6-volt dry pack

Circuit Notes

The circuit is a two-stage, direct-coupled transistor amplifier connected as a free-running multivibrator. Both the flash duration and flash interval can be changed by turning the potentiometer, R1.

NEON TUBE FLASHER



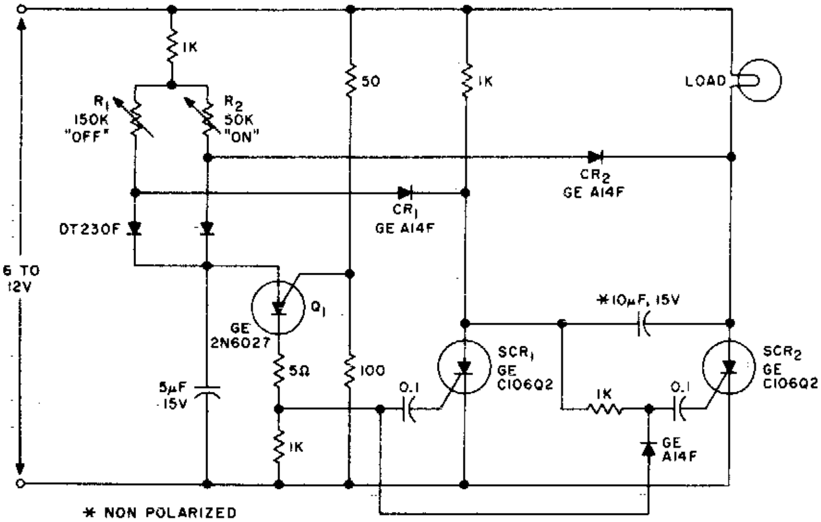
R2
10to 100k
(See text)
6.3v
240v
NEON
D1
1N4001
BY126
T1
h2
b1
e
Q1
2N2646
C2
470nF
Q2
BC108
Q3
BC108
R1
330Ω
+
C1
50µF
b1
e
b2
2N2646

Fig. 34-12

Circuit Notes

The voltage required to ignite the neon tube is obtained by using an ordinary filament transformer (240-6.3 V) in reverse. Battery drain is quite low, around 1 to 2 milliamps for a nine volt battery. The pulses from Q1, unijunction transistor, operated as a relaxation oscillator and are applied to Q2 which in turn drives Q3 into saturation. The sharp rise in current through the 6.3 V winding of the transformer as Q3 goes into saturation induces a high voltage in the secondary winding causing the neon to flash. The diode D1 protects the transistor from high voltage spikes generated when switching currents in the transformer.

DC FLASHER WITH ADJUSTABLE ON AND OFF TIME



1K
R_1
150K
"OFF"
R_2
50K
"ON"
50
1K
LOAD
CR_2
GE A14F
CR_1
GE A14F
DT230F
6 TO
12V
Q_1
GE
2N6027
$*10\mu F$, 15V
SCR_1
GE
C106Q2
SCR_2
GE
C106Q2
$5\mu F$
15V
5Ω
100
0.1
1K
0.1
GE
A14F
1K

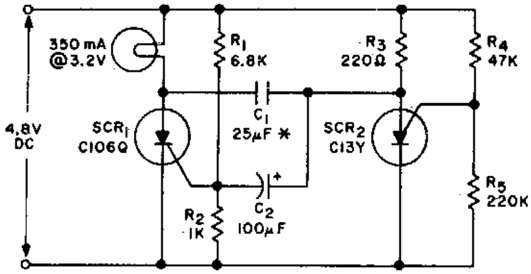
Fig. 34-13

* NON POLARIZED
ALL RESISTORS – 1/2 WATT
FLASH RATE AND DWELL TIME ADJUSTED BY R_1 AND R_2

Circuit Notes

This circuit utilizes a power flip-flop and programmable unijunction (PUT) to obtain adjustable on and off times.

LOW VOLTAGE FLASHER



350 mA
@3.2V
R_1
6.8K
R_3
220Ω
R_4
47K
4.8V
DC
SCR_1
C106Q
C_1
$25\mu F$ *
SCR_2
C13Y
+
R_5
220K
R_2
1K
C_2
$100\mu F$

Fig. 34-14

LAMP OFF • .5 – .6 SEC
LAMP ON ≈ 100 mSEC
* NON POLARIZED

Circuit Notes

Applying voltage to the circuit triggers SCR1. With SCR1 on, the voltage on the anode of SCR2 rises until SCR2 triggers to commutate SCR1. The voltage on the gate of SCR1 will swing negative at this time, and only after a positive potential of ≈ 0.5 volt is once again attained, will SCR1 retrigger. The circuit could be used for higher voltage levels, but the peak negative voltage on the gate of SCR1 must be limited to less than 6 volts.

1 A LAMP FLASHER

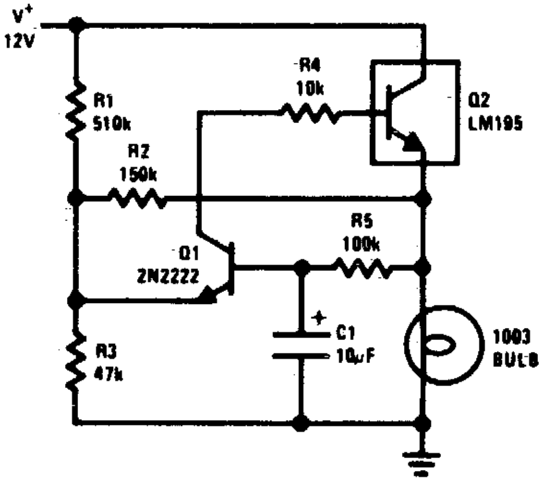
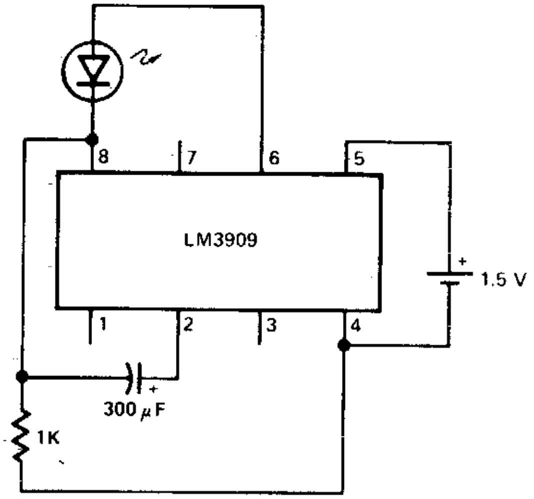

V+
12V
R1
510k
R2
150k
R4
10k
Q2
LM195
Q1
2N2222
R5
100k
+
C1
10µF
R3
47k
1003
BULB

Fig. 34-15

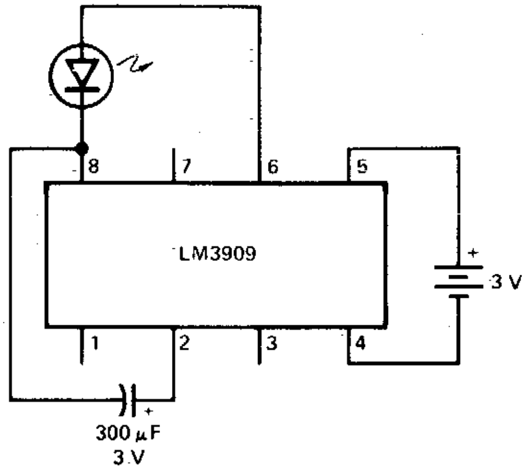
FAST BLINKER


8
7
6
5
LM3909
1
2
3
4
+
1.5 V
+
300 µF
1K

Note: Nominal Flash Rate:
2.6 Hz. Average I_{DRAIN} = 1.2

Fig. 34-16

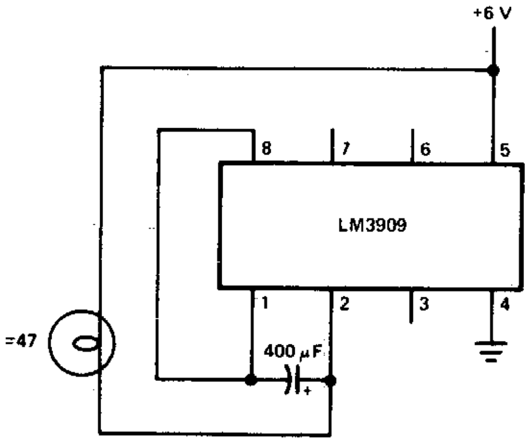
3 V FLASHER


8
7
6
5
LM3909
+
3 V
1
2
3
4
+
300 µF
3 V

Note: Nominal Flash Rate:
1 Hz. Average I_{DRAIN} = 0.77 mA

Fig. 34-17

INCANDESCENT BULB FLASHER


+6 V
8
7
6
5
LM3909
1
2
3
4
=47
400 µF
+

Note: Flash Rate: 1.5 Hz

Fig. 34-18

FLASHER FOR 4 PARALLEL LEDs

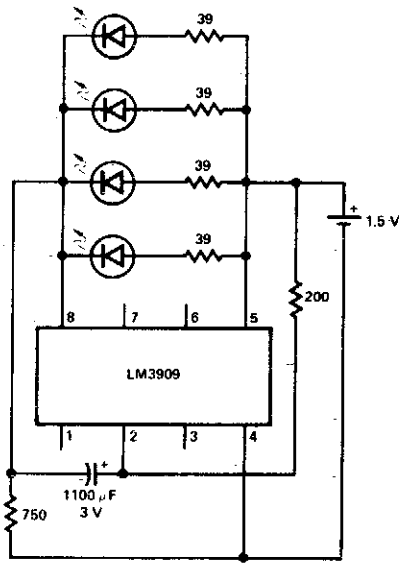
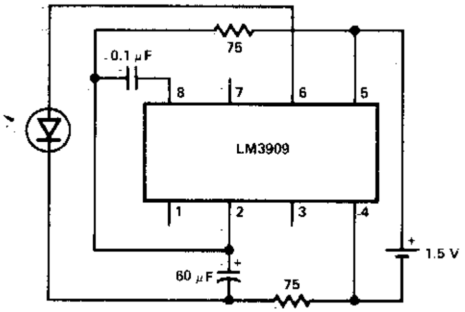

39
39
39
39
1.5 V
200
8
7
6
5
LM3909
1
2
3
4
1100 µF
3 V
750

Fig. 34-19

Note: Nominal Flash Rate:
1.3 Hz. Average I_{DRAIN} = 2 mA

LED BOOSTER


75
0.1 µF
8
7
6
5
LM3909
1
2
3
4
1.5 V
60 µF
75

Note: High efficiency, 4 mA drain

Note: Continuous Appearing Light Obtained By Supplying Short, High Current, Pulses (2 kHz) to LEDs With Higher Than Battery Voltage Available.

Fig. 34-20

SAFE, HIGH VOLTAGE FLASHER

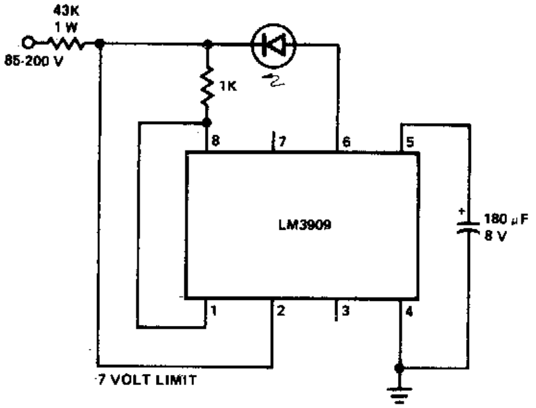

43K
1 W
85-200 V
1K
8
7
6
5
LM3909
180 µF
8 V
1
2
3
4
7 VOLT LIMIT

Fig. 34-21

ALTERNATING FLASHER

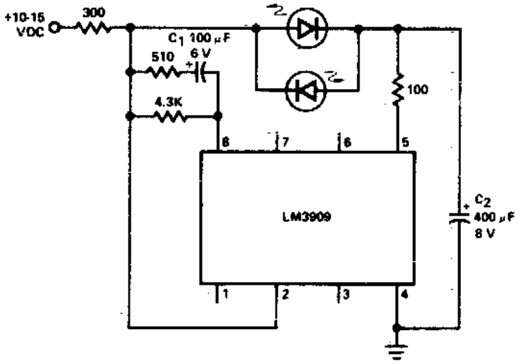
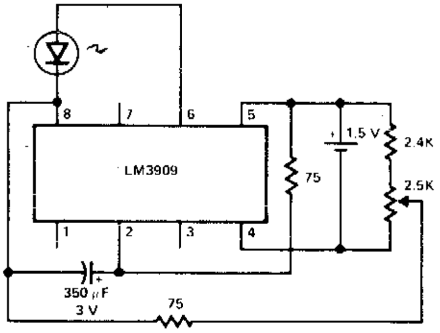

+10-15
VDC
300
C1 100 µF
6 V
510
4.3K
100
8
7
6
5
LM3909
C2
400 µF
8 V
1
2
3
4

Fig. 34-22

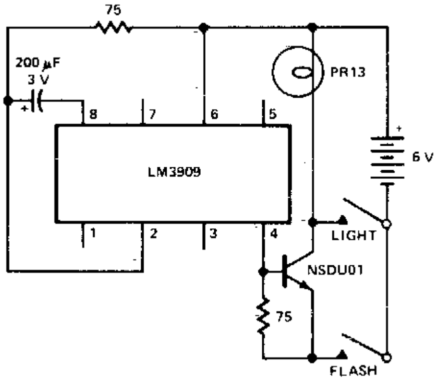
VARIABLE FLASHER


8
7
6
5
LM3909
1
2
3
4
1.5 V
2.4K
75
2.5K
350 μF
3 V
75

Note: Flash Rate: 0-20 Hz

Fig. 34-23

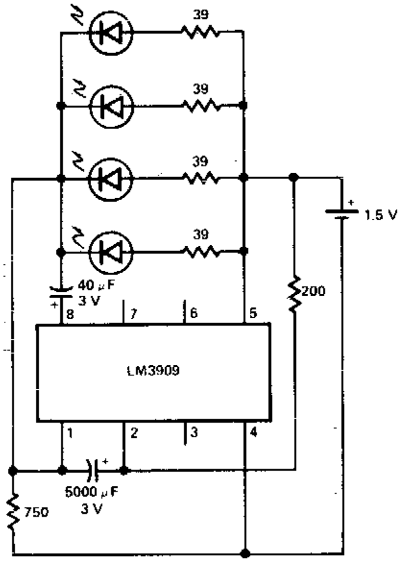
EMERGENCY LANTERN/FLASHER


75
200 μF
3 V
8
7
6
5
LM3909
1
2
3
4
PR13
6 V
LIGHT
NSDU01
75
FLASH

Note: Nominal Flash Rate: 1.5 Hz

Fig. 34-24

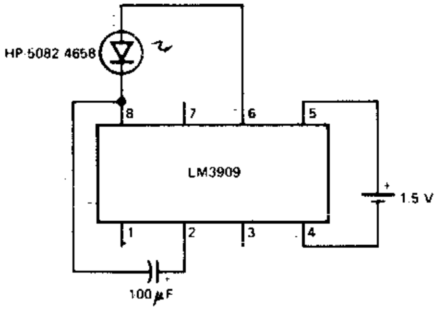
HIGH EFFICIENCY
PARALLEL CIRCUIT FLASHER


39
39
39
39
1.5 V
40 μF
3 V
200
8
7
6
5
LM3909
1
2
3
4
5000 μF
3 V
750

Note: Nominal Flash Rate:
1.5 Hz. Average I_{DRAIN} = 1.5 mA

Fig. 34-25

MINIMUM POWER FLASHER (1.5 V)


HP 5082 4658
8
7
6
5
LM3909
1.5 V
1
2
3
4
100 μF

Note: Nominal Flash Rate: 1.1 Hz. Average I_{DRAIN}
= 0.32 mA

Fig. 34-26

35

Frequency Measuring Circuits

The sources of the following circuits are contained in the Sources section beginning on page 730. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

Inexpensive Frequency Counter/Tachometer
Audio Frequency Meter
Linear Frequency Meter
Power-Line Frequency Meter

INEXPENSIVE FREQUENCY COUNTER/TACHOMETER

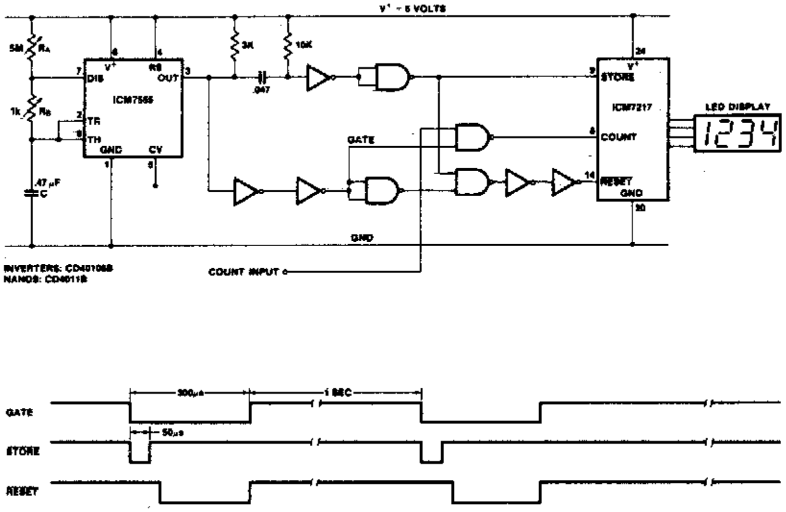

V⁺ = 5 VOLTS
5M
R_A
1k
R_B
.47 μF
C
ICM7555
DIS
TR
TH
GND
CV
V⁺
RS
OUT
3K
10K
.047
GATE
GND
ICM7217
STORE
COUNT
RESET
GND
LED DISPLAY
1234
INVERTERS: CD40106B
NANDS: CD4011B
COUNT INPUT
GATE
STORE
RESET
300μs
1 SEC
50μs

Fig. 35-1

Circuit Notes

This circuit uses the low power ICM7555 (CMOS 555) to generate the gating, STORE and RESET signals. To provide the gating signal, the timer is configured as an astable multivibrator. The system is calibrated by using a 5 M potentiometer for R_A as a coarse control and a 1 jk potentiometer for R_B as a fine control. CD40106B's are used as a monostable multivibrator and reset time delay.

LINEAR FREQUENCY METER (AUDIO SPECTRUM)

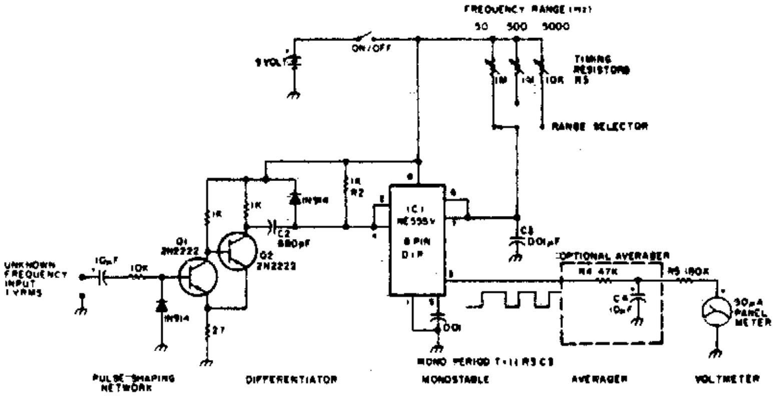

FREQUENCY RANGE (HZ)
50 500 5000
9 VOLT
ON/OFF
TIMING RESISTORS R5
1M
1M
10K
RANGE SELECTOR
1K R2
IN914
1K
1K
C2 680pF
Q1 2N2222
Q2 2N2222
IC1 NE555V 8 PIN DIP
C3 .001μF
UNKNOWN FREQUENCY INPUT 1 VRMS
10μF
10K
IN914
27
.001
OPTIONAL AVERAGER
R4 47K
R5 180K
C4 10μF
50μA PANEL METER
MONO PERIOD T=1.1 R5 C3
PULSE SHAPING NETWORK
DIFFERENTIATOR
MONOSTABLE
AVERAGER
VOLTMETER

Fig. 35-2

Circuit Notes

The 555 is used in a monostable multivibrator circuit that puts out a fixed timewidth pulse, which is triggered by the unknown input frequency.

POWER-LINE FREQUENCY METER

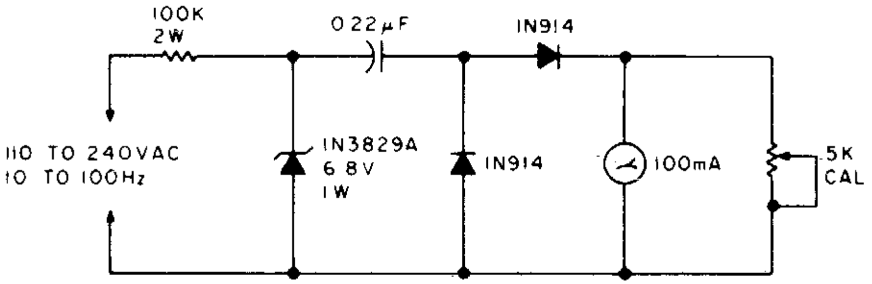

100K
2W
0.22μF
IN914
110 TO 240VAC
10 TO 100Hz
IN3829A
6.8V
1W
IN914
100mA
5K
CAL

Fig. 35-3

Circuit Notes

The meter will indicate the frequency from a power generator. Incoming sine waves are converted to square waves by the 100 K resistor and the 6.8 V zener. The square wave is differentiated by the capacitor and the current is averaged by the diodes. The average current is almost exactly proportional to the frequency and can be read directly on a 100 mA meter. To calibrate, hook the circuit up to a 60 Hz poweraline and adjust the 5 K pot to read 60 mA.

AUDIO FREQUENCY METER

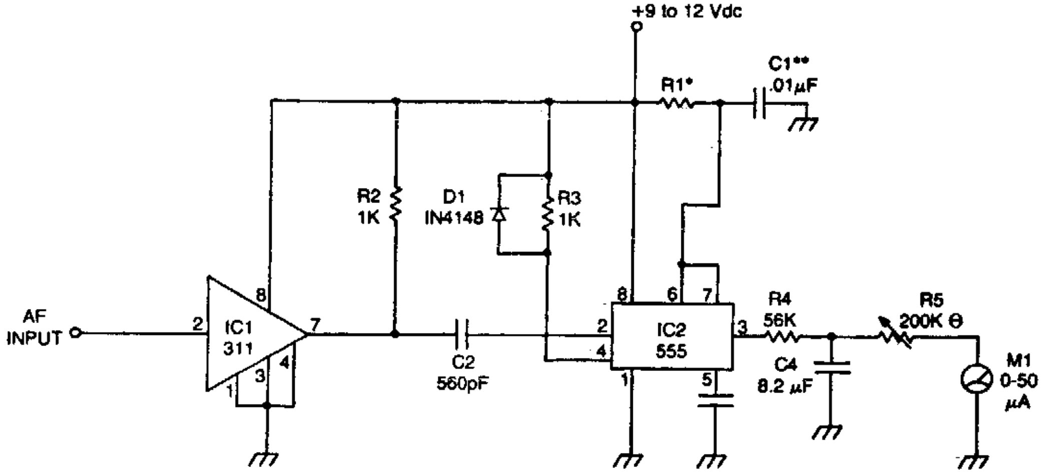

+9 to 12 Vdc
R1*
C1**
.01μF
R2
1K
D1
IN4148
R3
1K
AF
INPUT
IC1
311
8
2
7
3
4
1
C2
560pF
IC2
555
8
6
7
2
4
3
1
5
R4
56K
C4
8.2 μF
R5
200K Θ
M1
0-50
μA

Fig. 35-4

*R1: 1 meg 100K 10K
RANGE: 50 500 5000
(Hz)
**C1 : POLYSTYRENE OR SILVER-MICA

Circuit Notes

The meter uses time averaging to produce a direct current that is proportional to the frequency of the input signal.

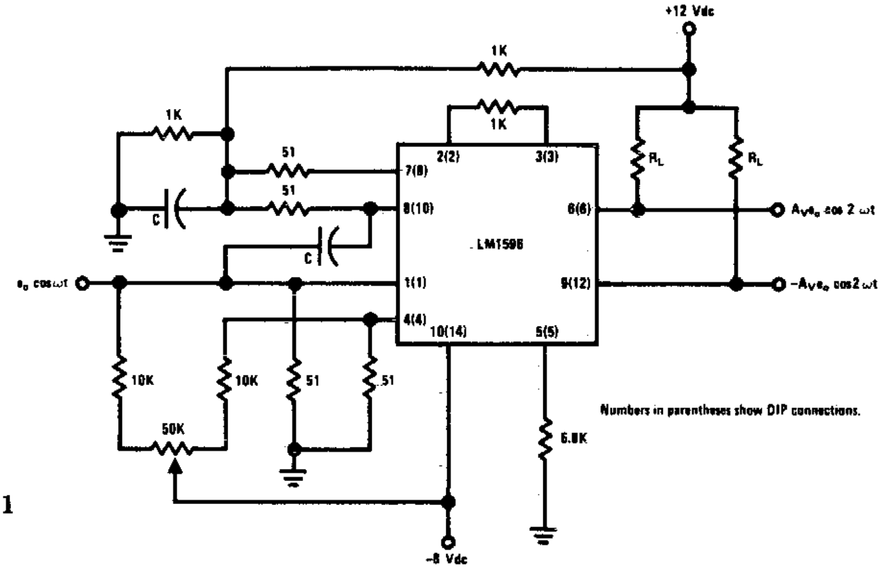
36

Frequency Multipliers

The sources of the following circuits are contained in the Sources section beginning on page 730. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

Broadband Frequency Doubler
Frequency Doubler
150 to 300 MHz Doubler
Low-Frequency Doubler
Oscillator with Double Frequency Output

BROADBAND FREQUENCY DOUBLER


+12 Vdc
1K
1K
1K
51
51
2(2)
3(3)
7(8)
8(10)
6(6)
R_L
R_L
C
C
LM1596
$e_o \cos \omega t$
1(1)
9(12)
4(4)
10(14)
5(5)
$A_V e_o \cos 2\omega t$
$-A_V e_o \cos 2\omega t$
10K
10K
51
51
50K
6.8K
-8 Vdc

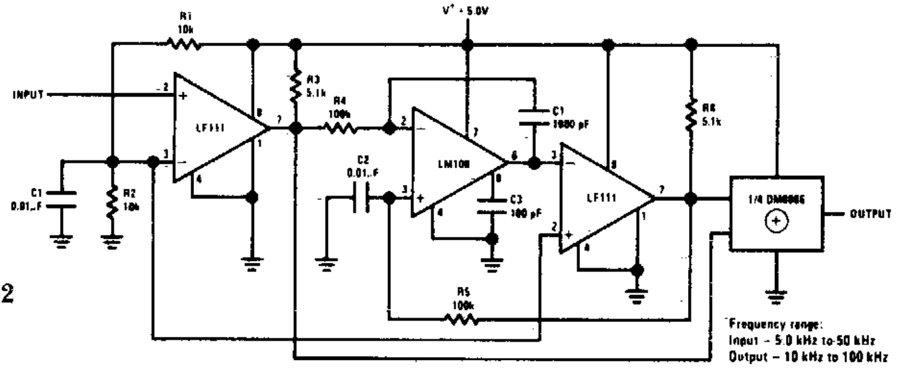
Numbers in parentheses show DIP connections.

Fig. 36-1

Circuit Notes

This circuit will double low-level signals with low distortion. The value of C should be chosen for low reactance at the operating frequency. Signal level at the carrier input must be less than 25 mV peak to maintain operation in the linear region of the switching differential amplifier. Levels to 50 mV peak may be used with some distortion of the output waveform. If a larger input signal is available, a resistive divider may be used at the carrier input with full signal applied to the signal input.

FREQUENCY DOUBLER


R1
10k
$V^+ = 5.0V$
INPUT
2
3
LF111
8
7
1
4
R3
5.1k
R4
100k
C1
0.01μF
R2
10k
C2
0.01μF
LM108
C1
1000 pF
C3
100 pF
LF111
R6
5.1k
1/4 DM8086
OUTPUT
R5
100k

Frequency range:
Input – 5.0 kHz to 50 kHz
Output – 10 kHz to 100 kHz

Fig. 36-2

150 TO 300 MHz DOUBLER

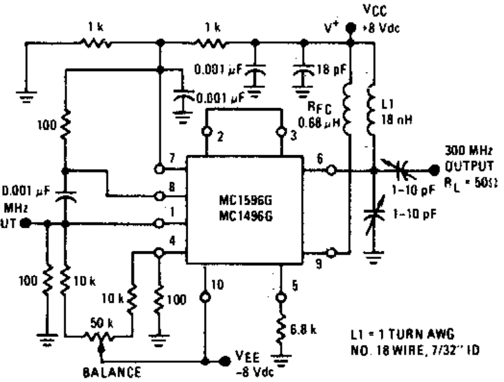

VCC
+8 Vdc
V+
1 k
1 k
0.001 μF
0.001 μF
18 pF
RFC
0.68 μH
L1
18 nH
100
2
3
7
6
8
1
4
9
10
5
MC1596G
MC1496G
0.001 μF
150 MHz
INPUT
300 MHz
OUTPUT
RL = 50Ω
1–10 pF
1–10 pF
100
10 k
10 k
100
50 k
6.8 k
BALANCE
VEE
−8 Vdc
L1 = 1 TURN AWG
NO. 18 WIRE, 7/32" ID

Fig. 36-3

LOW-FREQUENCY DOUBLER

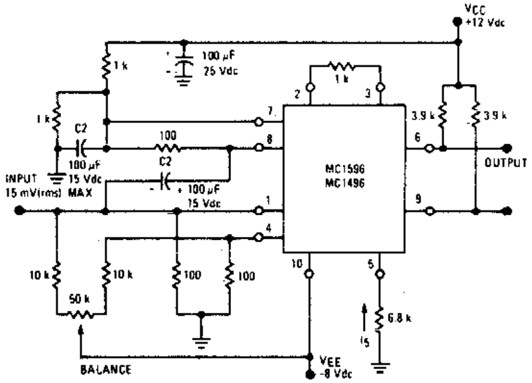

VCC
+12 Vdc
100 μF
25 Vdc
1 k
1 k
1 k
2
3
7
8
1
4
6
9
10
5
C2
100 μF
15 Vdc
100
C2
100 μF
15 Vdc
INPUT
15 mV(rms) MAX
MC1596
MC1496
3.9 k
3.9 k
OUTPUT
10 k
10 k
100
100
50 k
BALANCE
6.8 k
I5
VEE
−8 Vdc

Fig. 36-4

OSCILLATOR WITH DOUBLE FREQUENCY OUTPUT

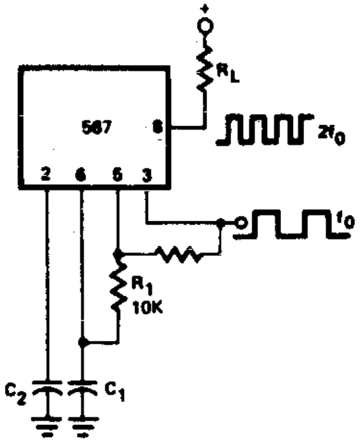

+
RL
567
8
2
6
5
3
2f0
f0
R1
10K
C2
C1

Fig. 36-5

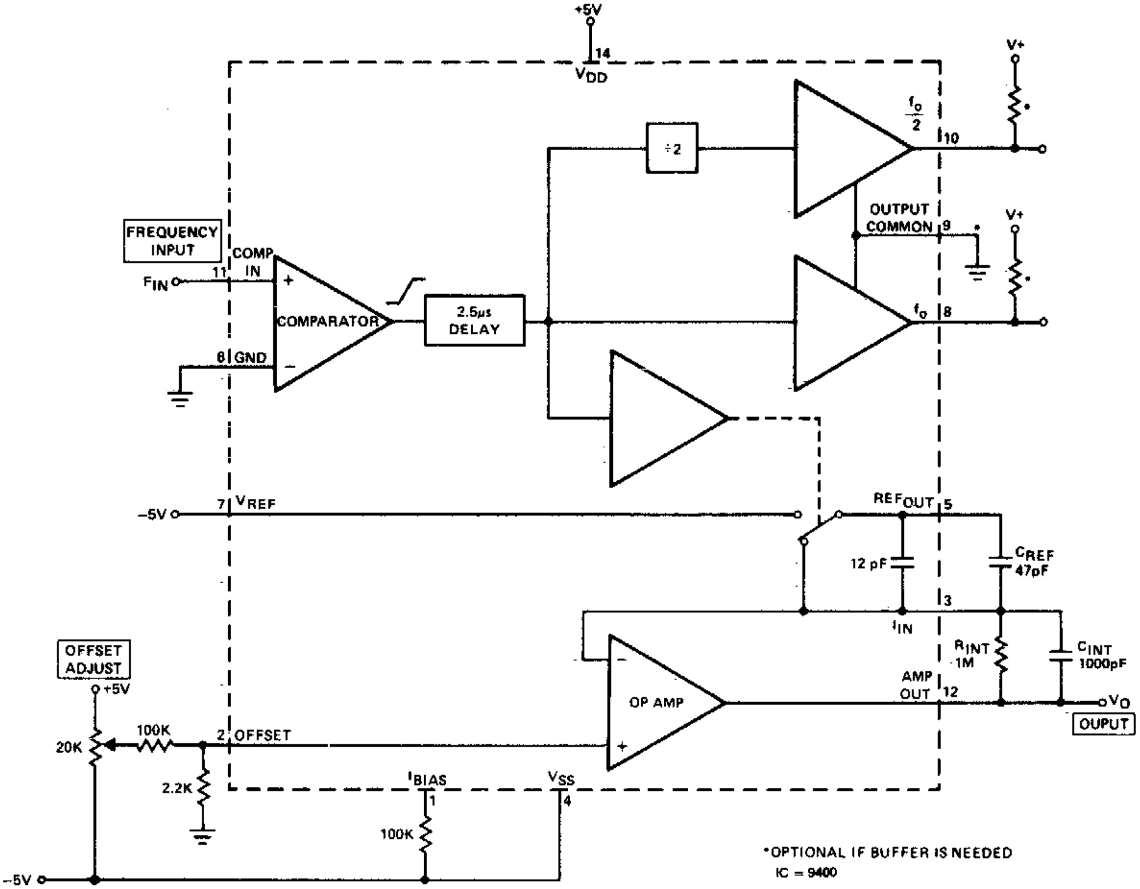
37

Frequency-to-Voltage Converters

The sources of the following circuits are contained in the Sources section beginning on page 730. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

DC-10 kHz Frequency/Voltage Converter
Frequency-to-Voltage Converter
Zener Regulated Frequency-to-Voltage Converter
Simple Frequency-to-Voltage Converter
F/V Conversion, TTL Input
Frequency-to-Voltage Converter with 2-Pole Butterworth Filter to Reduce Ripple
Precision Frequency-to-Voltage Converter

DC-10 kHz FREQUENCY/VOLTAGE CONVERTER


+5V
14
V_{DD}
V+
$\frac{f_o}{2}$
10
÷2
OUTPUT COMMON
9
V+
FREQUENCY INPUT
COMP IN
11
F_{IN}
+
COMPARATOR
2.5μs DELAY
f_o
8
6 GND
−
−5V
7 V_{REF}
REF_{OUT} 5
12 pF
C_{REF} 47pF
3
I_{IN}
R_{INT} 1M
C_{INT} 1000pF
AMP OUT 12
V_O
OUPUT
OFFSET ADJUST
+5V
20K
100K
2 OFFSET
2.2K
OP AMP
I_{BIAS}
1
V_{SS}
4
100K
−5V

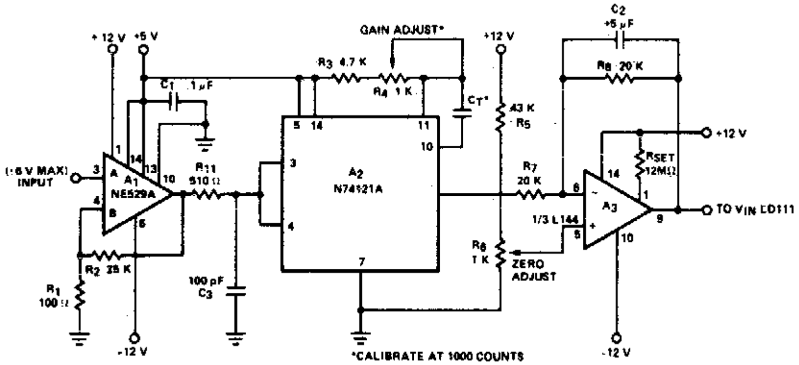
*OPTIONAL IF BUFFER IS NEEDED
IC = 9400

Fig. 37-1

Circuit Notes

The converter generates an output voltage which is linearly proportional to the input frequency waveform. Each zero crossing at the comparator's input causes a precise amount of change to be dispensed into the op amp's summing junction. This charge in turn flows through the feedback resistor generating voltage pulses at the output of the op amp. Capacitor (CINT) across RINT averages these pulses into a dc voltage which is linearly proportional to the input frequency.

FREQUENCY-TO-VOLTAGE CONVERTER (DIGITAL FREQUENCY METER)


+12 V
+5 V
C1 .1 μF
(±6 V MAX) INPUT
3
4
1
14
13
10
A
B
A1
NE529A
6
R2 25 K
R1 100 Ω
-12 V
R11 510 Ω
100 pF C3
GAIN ADJUST*
R3 4.7 K
R4 1 K
CT*
5
14
11
10
3
4
7
A2
N74121A
+12 V
43 K R5
R7 20 K
R6 1 K
ZERO ADJUST
C2 +5 μF
R8 20 K
+12 V
RSET 12MΩ
14
6
1
A3
1/3 L144
5
9
10
TO VIN LD111
-12 V
*CALIBRATE AT 1000 COUNTS

RANGE	C_T
2 kHz	.082 μF
20 kHz	.0082 μF
200 kHz	820 pF
2 MHz	82 pF
20 MHz	8.2 pF

Fig. 37-2

Circuit Notes

This circuit converts frequency to voltage by taking the average dc value of the pulses from the 74121 monostable multivibrator. The one shot is triggered by the positive-going ac signal at the input of the 529 comparator. The amplifier acts as a dc filter, and also provides zeroing. The accuracy is 2% over a 5 decade range. The input signal to the comparator should be greater than 0.1 volt peak-to-peak, and less than 12 volts peak-to-peak for proper operation.

ZENER REGULATED FREQUENCY-TO-VOLTAGE CONVERTER

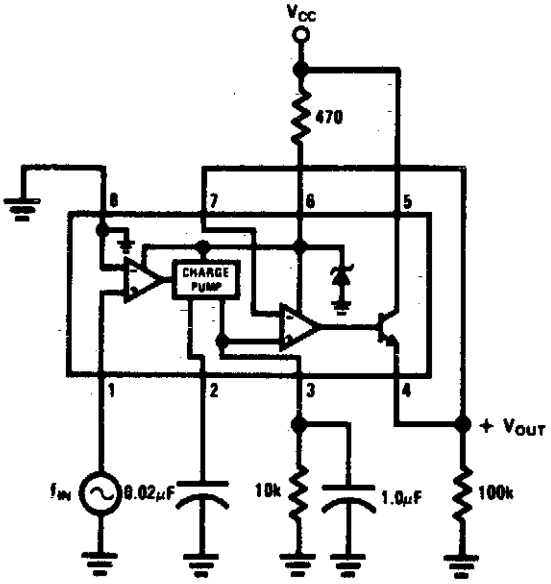
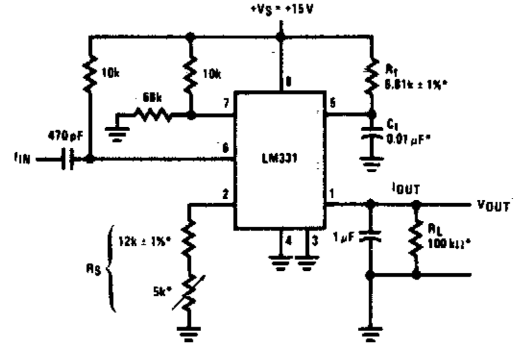

Vcc
470
8
7
6
5
CHARGE PUMP
1
2
3
4
+ VOUT
fIN
0.02μF
10k
1.0μF
100k

Fig. 37-3

SIMPLE FREQUENCY-TO-VOLTAGE CONVERTER (10 kHz FULL-SCALE, ±0.006% NON-LINEARITY)

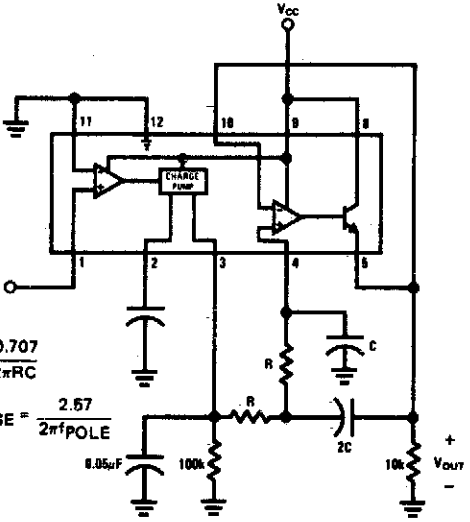

+VS = +15V
10k
10k
68k
470pF
fIN
8
7
6
2
5
1
4
3
LM331
Rt
6.81k ± 1%*
Ct
0.01 μF*
IOUT
VOUT
1 μF
RL
100 kΩ*
12k ± 1%*
RS
5k*

$V_{OUT} = f_{IN} \times 2.09V \times \frac{R_L}{R_S} \times (R_tC_t)$

*Use stable components with low temperature coefficients.

Fig. 37-4

FREQUENCY-TO-VOLTAGE CONVERTER WITH 2-POLE BUTTERWORTH FILTER TO REDUCE RIPPLE


VCC
11
12
10
9
8
CHARGE PUMP
1
2
3
4
5
fIN
C
R
R
2C
0.05μF
100k
10k
+
VOUT
−

$f_{POLE} = \frac{0.707}{2\pi RC}$

$\tau_{RESPONSE} = \frac{2.57}{2\pi f_{POLE}}$

Fig. 37-6

F/V CONVERSION, TTL INPUT

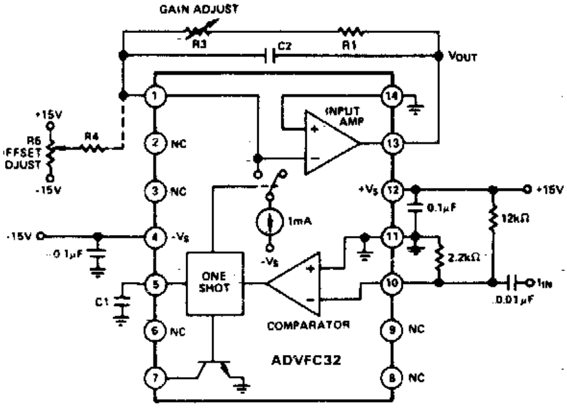
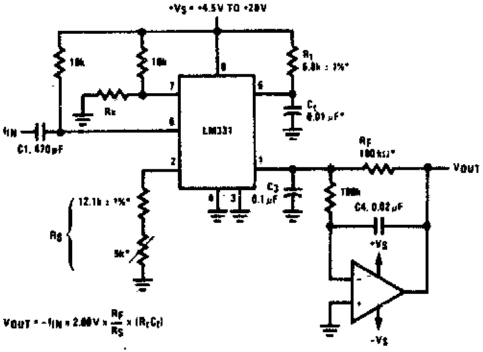

GAIN ADJUST
R3
C2
R1
VOUT
+15V
R5
OFFSET ADJUST
-15V
R4
1
2 NC
3 NC
4 -VS
5
6 NC
7
14
13
12 +VS
11
10
9 NC
8 NC
INPUT AMP
1mA
-VS
ONE SHOT
COMPARATOR
ADVFC32
-15V
0.1μF
C1
+15V
0.1μF
12kΩ
2.2kΩ
fIN
0.01μF

Fig. 37-5

PRECISION FREQUENCY-TO-VOLTAGE CONVERTER (10 kHz FULL/SCALE WITH 2-POLE FILTER, ±0.01% NON-LINEARITY MAXIMUM)


+VS = +4.5V TO +20V
10k
10k
Rx
fIN
C1, 470pF
8
7
6
2
5
1
4
3
LM331
Rt
6.8k ± 1%*
Ct
0.01 μF*
RF
100 kΩ*
VOUT
C3
0.1 μF
100k
C4, 0.02 μF
+VS
-VS
12.1k ± 1%*
RS
5k*

$V_{OUT} = -f_{IN} \times 2.09V \times \frac{R_F}{R_S} \times (R_tC_t)$

$\text{SELECT } R_X = \frac{(V_S - 2V)}{0.2mA}$

*Use stable components with low temperature coefficients.

Fig. 37-7

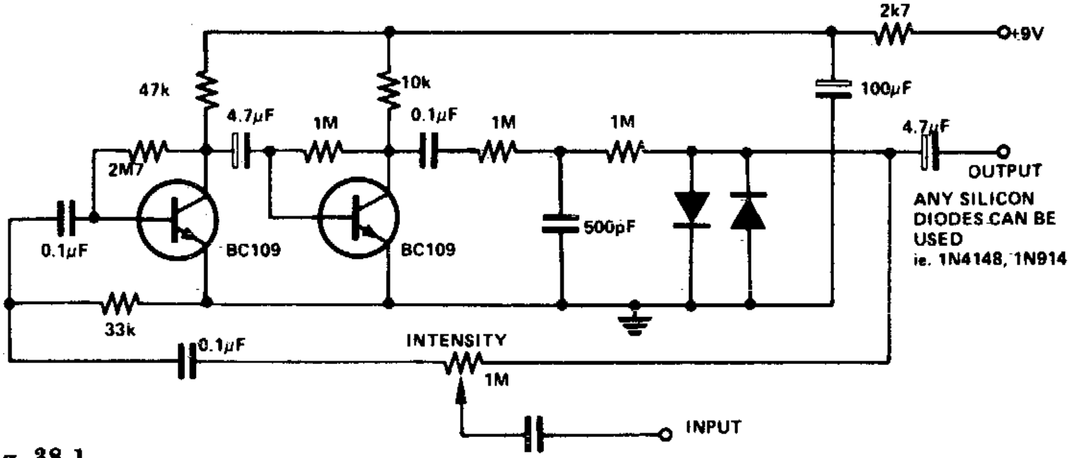
38

Fuzz Circuits

The sources of the following circuits are contained in the Sources section beginning on page 730. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

Fuzz Box 1
Fuzz Box 2
Fuzz Box 3
Fuzz Box 4
Fuzz Box 5
Guitar Fuzz

FUZZ BOX 1


2k7
+9V
47k
10k
4.7µF
1M
0.1µF
1M
1M
100µF
4.7µF
OUTPUT
2M7
0.1µF
BC109
BC109
500pF
33k
0.1µF
INTENSITY
1M
INPUT

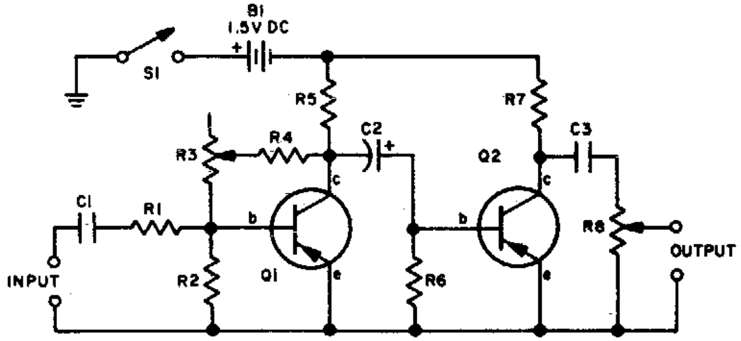
ANY SILICON DIODES CAN BE USED
ie. 1N4148, 1N914

Fig. 38-1

Circuit Notes

The input signal is amplified by the transistors. The distorted output is then clipped by the two diodes and the high frequency noise is filtered from the circuit via the 500 pF capacitor. The 1 M pot adjusts the intensity of the fuzz from maximum to no fuzz (normal playing).

FUZZ BOX 2


B1
1.5V DC
S1
R5
R7
C2
C3
R4
R3
Q2
C1
R1
b
c
e
b
c
e
R8
OUTPUT
INPUT
R2
Q1
R6

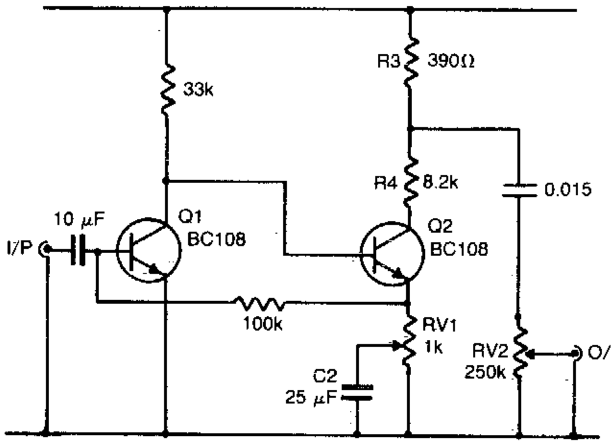
B1—1.5-V AA battery
C1, C3—0.1-uF, 50-VDC capacitor
C2—4.7-uF, 10-VDC electrolytic capacitor
Q1, Q2—pnp transistor—HEP-632
R1, R6—22,000-ohm, ½-watt resistor
R2—18,000-ohm, ½-watt resistor
R3—1-megohm pot
R4—100,000-ohm, ½-watt resistor
R5, R7—10,000-ohm, ½-watt resistor
R8—50,000-ohm pot
S1—Spst switch

Fig. 38-2

Circuit Notes

Potentiometer R3 sets the degree of fuzz, and R8 sets the output level. Since the fuzz effect cannot be completely eliminated by R3, fuzz-free sound requires a bypass switch from the input to output terminals.

FUZZ BOX 3



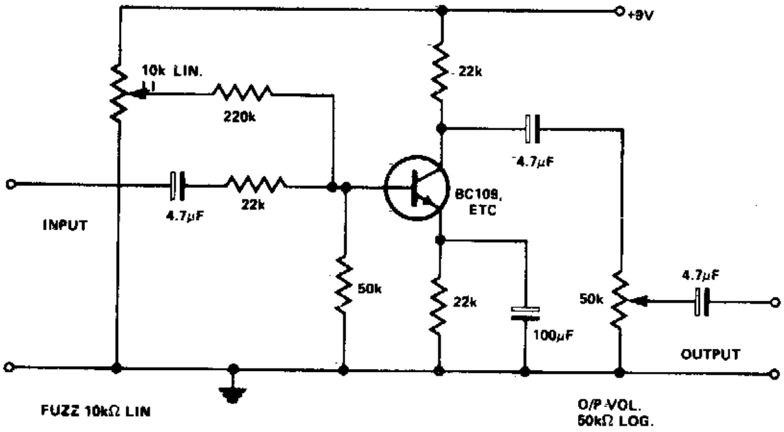
R3
390Ω
33k
R4
8.2k
0.015
10 μF
I/P
Q1
BC108
Q2
BC108
100k
RV1
1k
RV2
250k
O/P
C2
25 μF

Fig. 38-3

Circuit Notes

Q1 and Q2 form a voltage amplifier which has sufficient gain to be overdriven by a relatively low input, such as an electric guitar. The result is that the output from Q2 is a Squared-Off verson of the input, giving the required fuzz sound. RV1 adjusts the amount of negative feedback inserted into the circuit by C2, and thus the amount of squaring of the signal. The purpose of R3 and R4 is to lower the output voltage to a suitable level, which is then adjusted as required with the volume control VR2.

FUZZ BOX 4



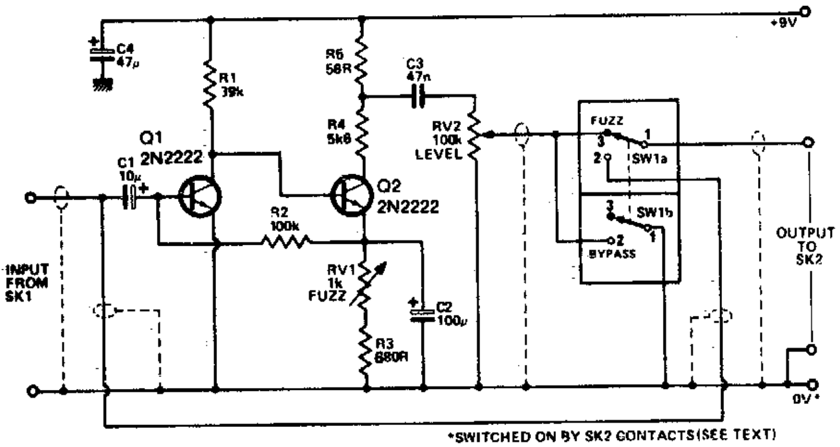
+9V
10k LIN.
220k
22k
4.7μF
INPUT
4.7μF
22k
BC108, ETC
50k
22k
100μF
50k
4.7μF
OUTPUT
FUZZ 10kΩ LIN
O/P VOL. 50kΩ LOG.

Fig. 38-4

Circuit Notes

None of the components are particularly critical in value or quality, as distortion is the sole object! The transistor could be BC107-8-9, 2N2926, etc.

FUZZ BOX 5


C4
47μ
R1
39k
Q1
2N2222
C1
10μ
INPUT
FROM
SK1
R6
56R
R4
5k6
R2
100k
Q2
2N2222
RV1
1k
FUZZ
R3
680R
C2
100μ
C3
47n
RV2
100k
LEVEL
FUZZ
3
1
2
SW1a
3
SW1b
2
1
BYPASS
+9V
OUTPUT
TO
SK2
0V*

*SWITCHED ON BY SK2 CONTACTS (SEE TEXT)

Circuit Diagram

Fig. 38-5

Circuit Notes

Transistors Q1 and Q2 amplify the incoming signal, and the gain is such that the input will overload when used with an electric guitar. RV1 adjusts the amount of feedback present, and hence voltage gain. The output is, therefore, a squared version of the input signal. The amount of squaring is varied by RV1.

GUITAR FUZZ

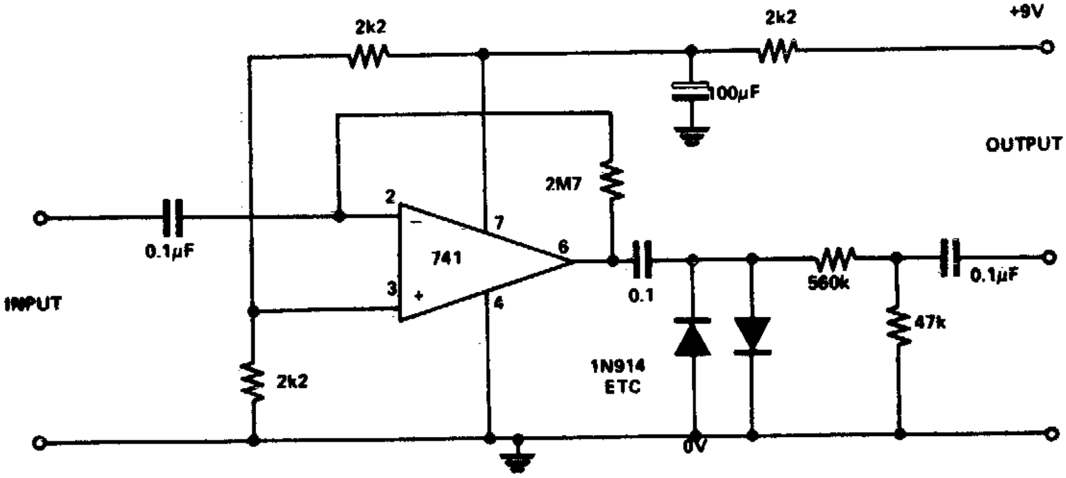

2k2
2k2
+9V
100μF
OUTPUT
2M7
2
7
741
6
3
4
0.1μF
INPUT
0.1
560k
0.1μF
47k
2k2
1N914
ETC
0V

Fig. 38-6

Circuit Notes

The 741 has a maximum gain of 20,000, but the circuit is so designed that the IC's gain is 2,700,000 which then distorts the output. This distortion gives the fuzz effect. The two diodes clip the output to drop the level, also lowered by the potential divider. This circuit also sustains the notes, due to clipping, giving a totally new sound.

39

Games

The sources of the following circuits are contained in the Sources section beginning on page 730. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

- Ready, Set, Go!
- Electronic Dice
- Game Roller or Chase Circuit
- Toss-A-Coin Binary Box
- Electronic Coin Tosser
- Heads or Tails
- Pot Shot
- Low Cost Heads or Tails
- Who Is First
- Windicator

READY, SET, GO!

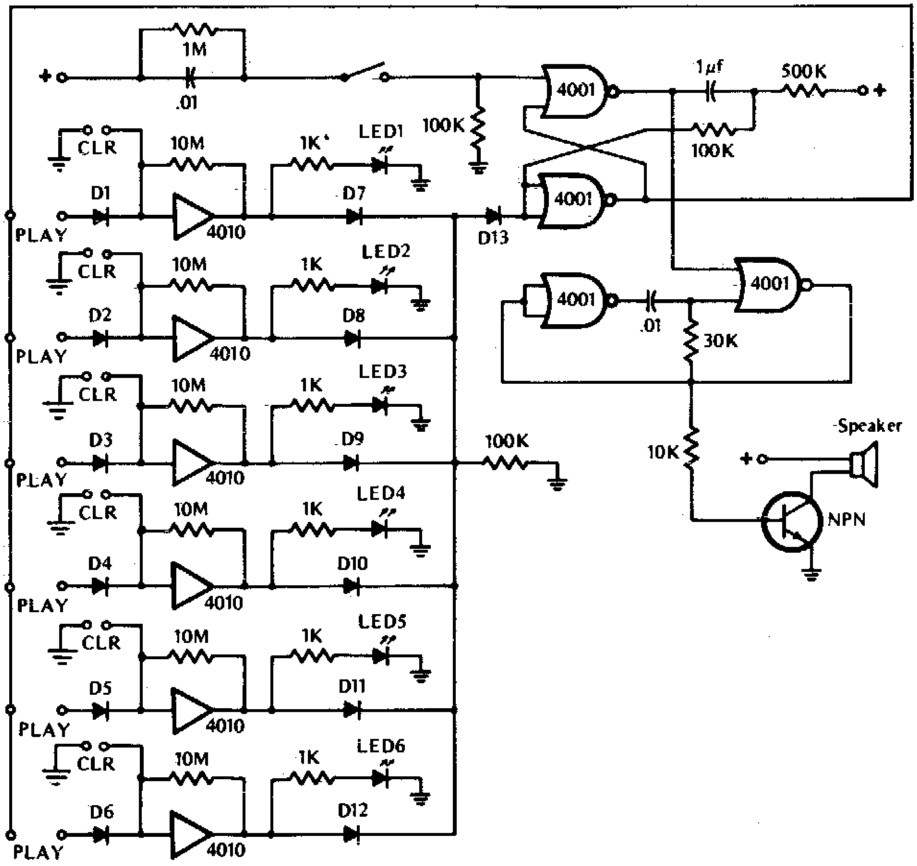
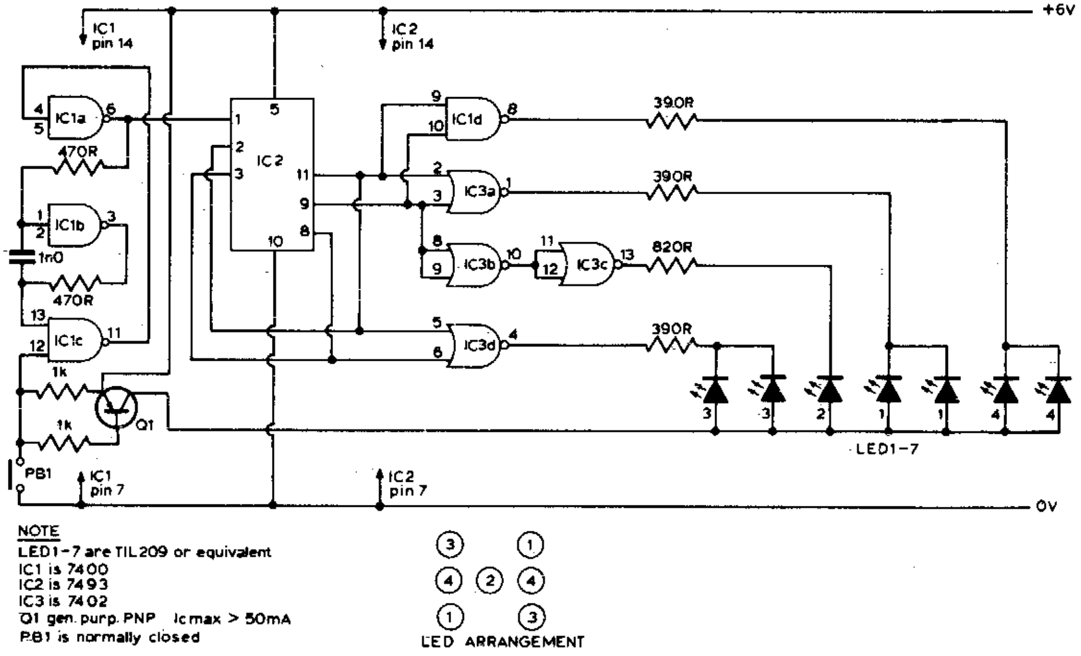

1M
.01
+
CLR
10M
1K
LED1
100K
D1
D7
PLAY
4010
D13
4001
1μf
500K
+
100K
LED2
D2
D8
.01
30K
LED3
D3
D9
100K
10K
Speaker
NPN
LED4
D4
D10
LED5
D5
D11
LED6
D6
D12

Fig. 39-1

Circuit Notes

This game tests a player's reaction time. It is activated by closing switch S1, which starts the tone generator and arms the circuit. The touchplate, labeled PLAY in the diagram, consists of two metal strips about 1/16th-inch apart. The first player to bridge the gap with his or her finger turns off the tone and lights the associated LED indicator. A second touchplate, labeled CLR in the diagram, clears the circuit, extinguishing the LED, when its gap is bridged by a fingertip.

ELECTRONIC DICE


+6V
IC1 pin 14
IC2 pin 14
IC1a
IC1b
IC1c
IC1d
470R
470R
1n0
1k
1k
Q1
PB1
IC1 pin 7
IC2 pin 7
IC2
IC3a
IC3b
IC3c
IC3d
390R
390R
820R
390R
LED1-7
0V
LED ARRANGEMENT

NOTE
LED1-7 are TIL209 or equivalent
IC1 is 7400
IC2 is 7493
IC3 is 7402
Q1 gen. purp. PNP Icmax > 50mA
PB1 is normally closed

Fig. 39-2

Circuit Notes

Six LEDs are arranged to produce a display the same as the dots on a dice. When PBI is depressed, the display is blanked and the oscillator (IC1 a, b, c) clocks IC2 at about 1MHz. IC2 counts from zero and resets on seven. When PBI is released, the display is enabled and a decoding system (IC3) produces the correct output on the LEDs.

GAME ROLLER OR CHASE CIRCUIT

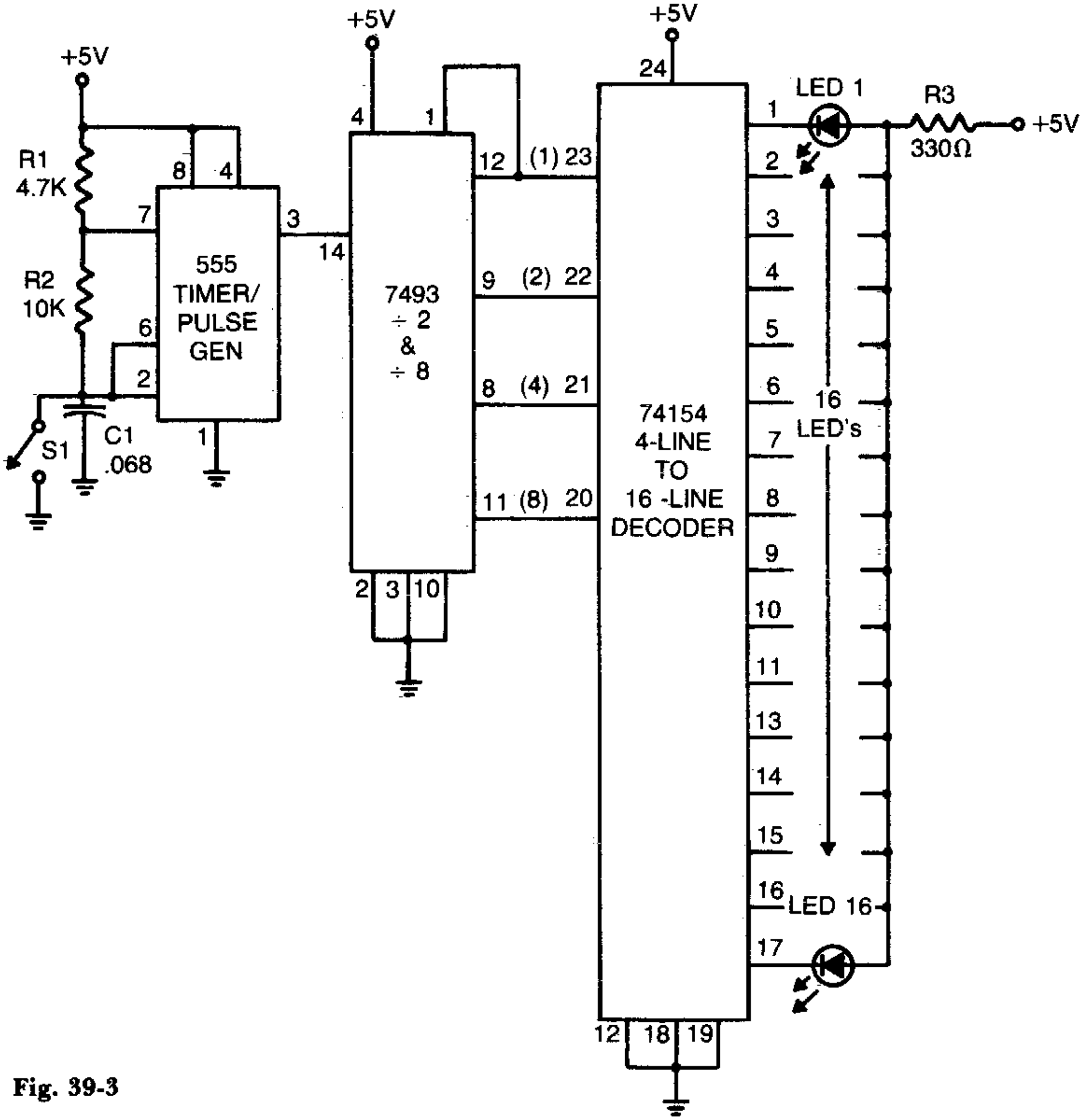
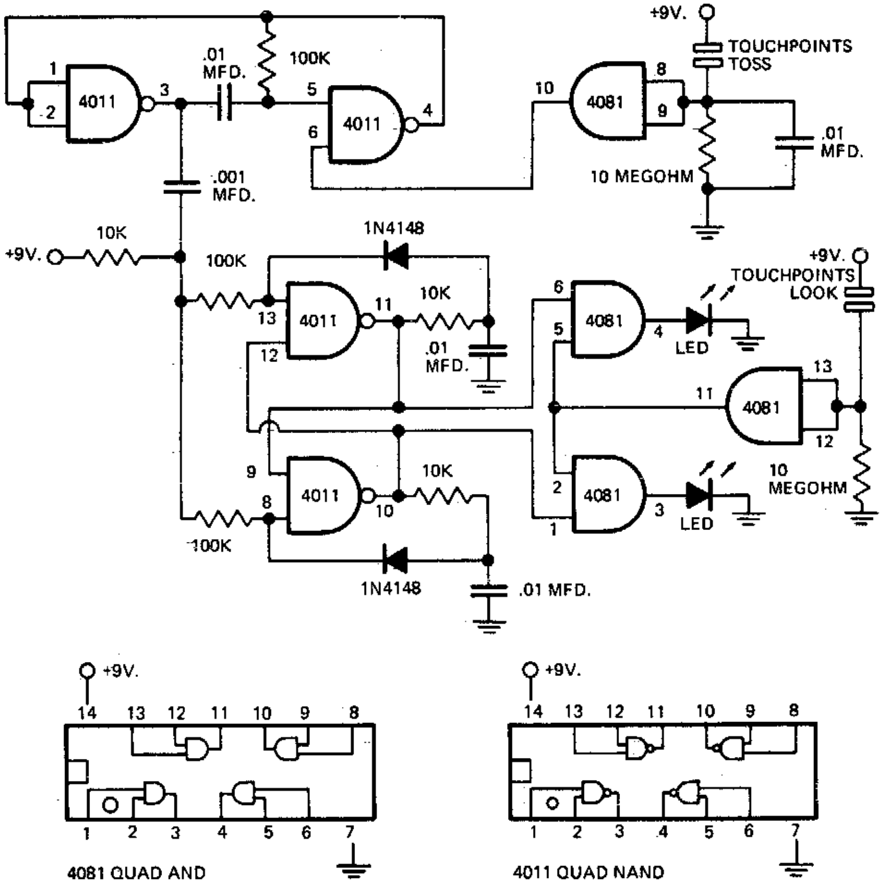

+5V
R1
4.7K
R2
10K
S1
C1
.068
8
4
7
6
2
1
3
555
TIMER/
PULSE
GEN
+5V
4
1
14
12
9
8
11
2
3
10
7493
÷ 2
&
÷ 8
(1) 23
(2) 22
(4) 21
(8) 20
+5V
24
74154
4-LINE
TO
16 -LINE
DECODER
12
18
19
1
2
3
4
5
6
7
8
9
10
11
13
14
15
16
17
LED 1
16
LED's
LED 16
R3
330Ω
+5V

Fig. 39-3

Circuit Notes

The 555 timer produces a rapid series of pulses whenever switch S1 is open. These pulses are counted in groups of 16 and converted into binary form by the 7493 and applied to the 74154 (a 1-of-16 decoder/demultiplexer) wired so that each of its 16 output lines goes low sequentially and in step with the binary count delivered by the 7493. When the switch is closed, only one LED remains on. Only one current limiting resistor (R3) is used for all the LED's since only one is on at any one time.

TOSS-A-COIN BINARY BOX



+9V.
TOUCHPOINTS
TOSS
4011
4011
4081
.01 MFD.
100K
10 MEGOHM
.01 MFD.
.001 MFD.
10K
+9V.
100K
1N4148
+9V.
TOUCHPOINTS
LOOK
4011
10K
.01 MFD.
4081
LED
4081
4081
LED
10 MEGOHM
4011
10K
100K
1N4148
.01 MFD.
+9V.
4081 QUAD AND
+9V.
4011 QUAD NAND

Fig. 39-4

Circuit Notes

Circuit uses an astable multivibrator to vary the heads-or-tails condition, and a flip-flop to store the condition given by the multivibrator. Consequently, the circuit is wired so that the flip-flop's state is changed once for each full cycle the multivibrator goes through to assure an absolutely even 50-50 chance of a heads or tails loss.

ELECTRONIC COIN TOSSER

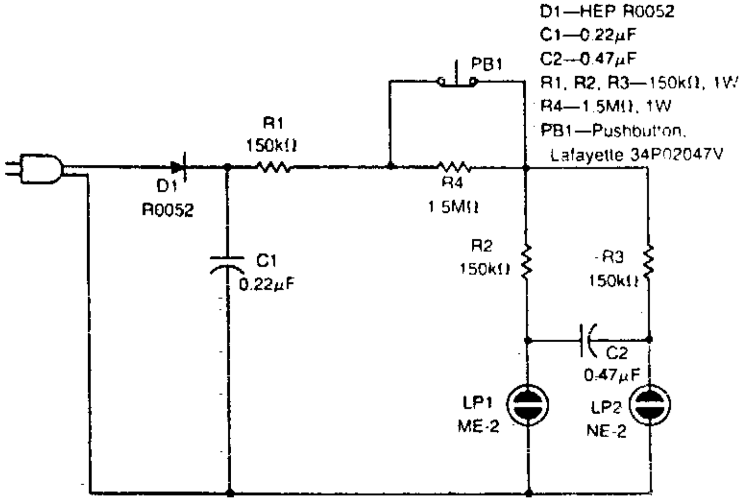

D1—HEP R0052
C1—0.22μF
C2—0.47μF
R1, R2, R3—150kΩ, 1W
R4—1.5MΩ, 1W
PB1—Pushbutton,
Lafayette 34P02047V
PB1
R1
150kΩ
D1
R0052
R4
1.5MΩ
C1
0.22μF
R2
150kΩ
R3
150kΩ
C2
0.47μF
LP1
ME-2
LP2
NE-2

Fig. 39-5

Circuit Notes

The circuit shown simulates the flipping of a coin by merely pushing switch PB1.

HEADS OR TAILS

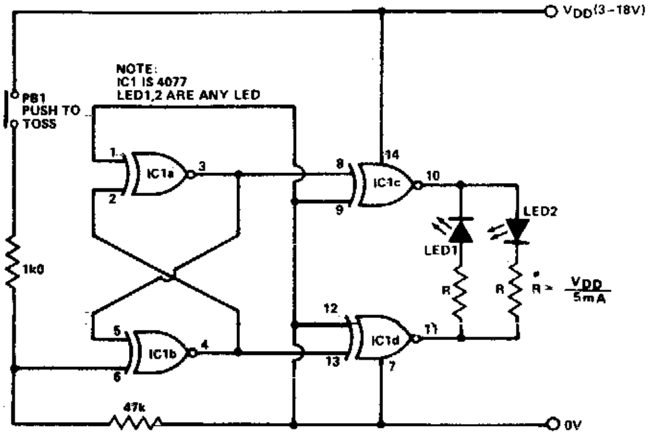

VDD(3-18V)
NOTE:
IC1 IS 4077
LED1,2 ARE ANY LED
PB1
PUSH TO
TOSS
1k0
47k
IC1a
IC1b
IC1c
IC1d
1
2
3
4
5
6
7
8
9
10
11
12
13
14
LED1
LED2
R
R
R = VDD/5mA
0V

Fig. 39-6

Circuit Notes

This ultra-simple heads or tails indicator uses a single 4077 and no capacitor.

The circuit is normally in a latched bistable mode; when the switch is closed the circuit will oscillate, i.e. toss the coin. The astable frequency is approximately 5-10 MHz. PB1 is a normally closed switch.

POT SHOT

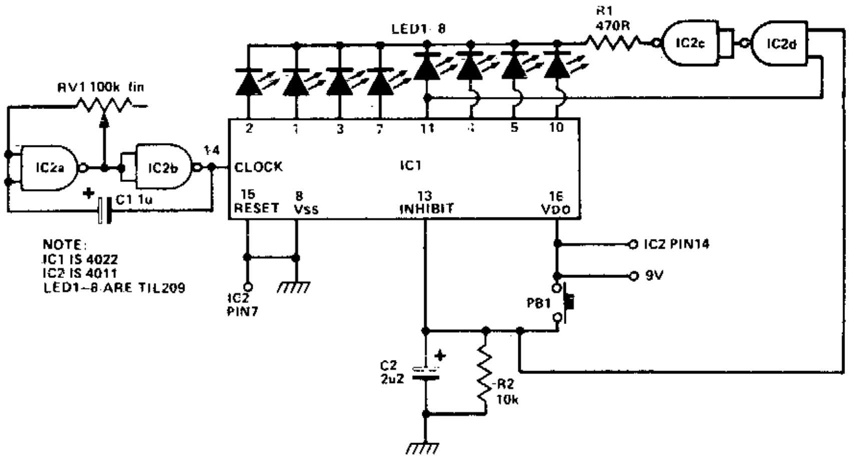

R1
470R
LED1 8
IC2c
IC2d
RV1 100k lin
2
1
3
7
11
4
5
10
IC2a
IC2b
14
CLOCK
IC1
15
RESET
8
Vss
13
INHIBIT
16
VDD
C1 1u
NOTE:
IC1 IS 4022
IC2 IS 4011
LED1–8 ARE TIL209
IC2
PIN7
IC2 PIN14
9V
PB1
C2
2u2
R2
10k

Fig. 39-7

Circuit Notes

This is a circuit for a game of the shooting gallery variety. IC2a and b form an astable multivibrator clocking IC1 which causes LEDs 1-8 to flash in turn LED 5 is the target LED and the object of the game is to depress PBI just as LED 5 comes on. If this is done, the whole display is blanked for a few seconds signifying a hit. Otherwise, the LED which was lit remains lit. When the push button is released, C2 discharges through R2 taking 8 pin 13 low again and the LEDs will start to flash again.

LOW COST "HEADS OR TAILS"

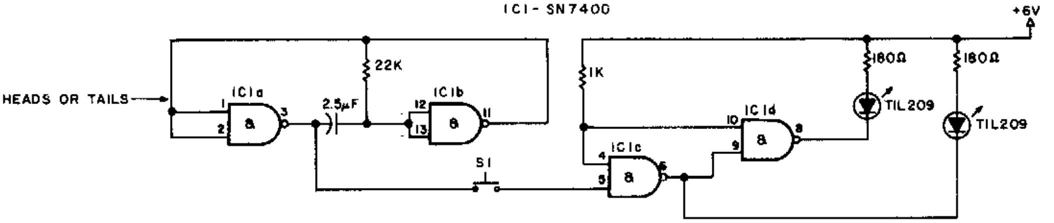

IC1 - SN7400
+6V
22K
1K
180Ω
180Ω
HEADS OR TAILS
IC1a
1
2
3
2.5μF
IC1b
12
13
11
IC1d
10
9
8
TIL209
TIL209
IC1c
4
5
6
S1

Fig. 39-8

Circuit Notes

S1 must be a push-to-make, release-to-break, switch.

WHO IS FIRST

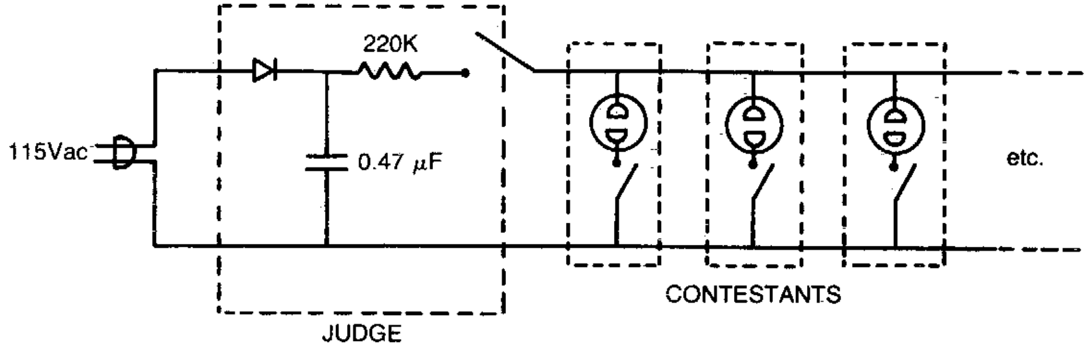

220K
115Vac
0.47 μF
etc.
CONTESTANTS
JUDGE

Fig. 39-9

Circuit Notes

Here is a circuit for any question-and-answer party game. The first button pushed unionizes the neon bulb dropping the dc voltage on the parallel neons (the other contestants) below the ionization level: determining unequivocally the first person to press the button.

WINDICATOR

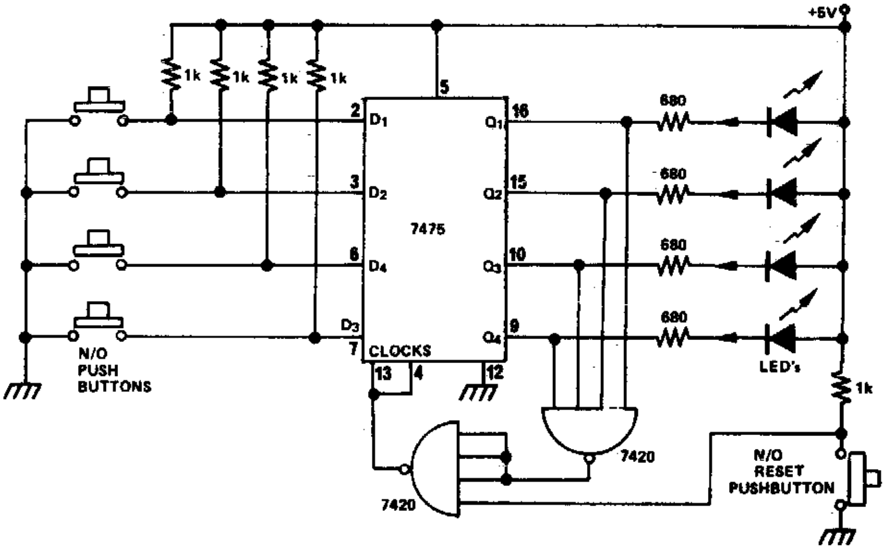

+5V
1k
1k
1k
1k
5
2
D1
Q1
16
3
D2
Q2
15
7475
6
D4
Q3
10
D3
Q4
9
7
CLOCKS
13
4
12
680
680
680
680
LED's
1k
N/O
PUSH
BUTTONS
7420
7420
N/O
RESET
PUSHBUTTON

Fig. 39-10

Circuit Notes

Two TTL ICs and a handful of other components are all that is needed for a circuit that will indicate which of four buttons was pressed first, as well as lock out all other entries. A logic 0 at one of the Q outputs, lights the appropriate LED and locks out other entries by taking the clock input low.

40

Gas/Vapor Detectors

The sources of the following circuits are contained in the Sources section beginning on page 730. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

Gas and Smoke Detector

Ionization Chamber Smoke Detector

Ionization Chamber Smoke Detector

GAS AND SMOKE DETECTOR

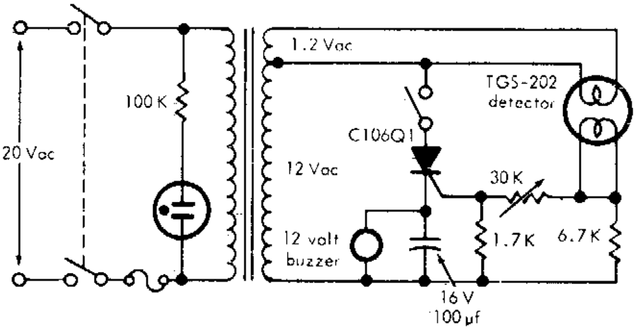

1.2 Vac
TGS-202
detector
100 K
C106Q1
120 Vac
12 Vac
30 K
12 volt
buzzer
1.7K
6.7K
16 V
100 μf

Fig. 40-1

Circuit Notes

This circuit can detect smoke and a number of gases (CO, CO_2, methane, coal gas and others) with a 10 ppm sensitivity. It uses a heated-surface semiconductor sensor. Detection occurs when the gas concentration increase causes a decrease of the sensor element internal resistance. The switch in series with the SCR is used for resetting the alarm.

IONIZATION CHAMBER SMOKE DETECTOR

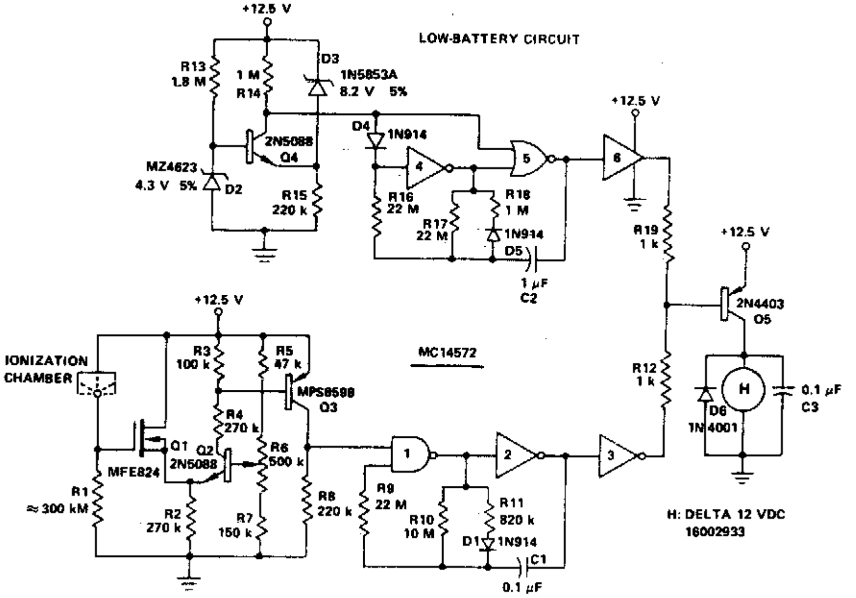

+12.5 V
LOW-BATTERY CIRCUIT
R13
1.8 M
1 M
R14
D3
1N5853A
8.2 V 5%
+12.5 V
D4
1N914
2N5088
Q4
MZ4623
4.3 V 5%
D2
R15
220 k
4
5
6
R16
22 M
R17
22 M
R18
1 M
1N914
D5
1 μF
C2
R19
1 k
+12.5 V
2N4403
Q5
+12.5 V
IONIZATION
CHAMBER
R3
100 k
R5
47 k
MPS8598
Q3
MC14572
R12
1 k
D6
1N4001
H
0.1 μF
C3
R4
270 k
Q1
Q2
2N5088
R6
500 k
MFE824
1
2
3
R1
≈ 300 kM
R2
270 k
R7
150 k
R8
220 k
R9
22 M
R10
10 M
R11
820 k
D1
1N914
C1
0.1 μF
H: DELTA 12 VDC
16002933

Fig. 40-2

Circuit Notes

Battery-operated, ionization chamber smoke detector includes a circuit to generate a unique alarm when the battery reaches the end of its useful life. The circuit uses the MCMOS MC14572 for two alarm oscillators (smoke and low battery). This circuit additionally uses five discrete transistors as buffers and comparators.

IONIZATION CHAMBER SMOKE DETECTOR

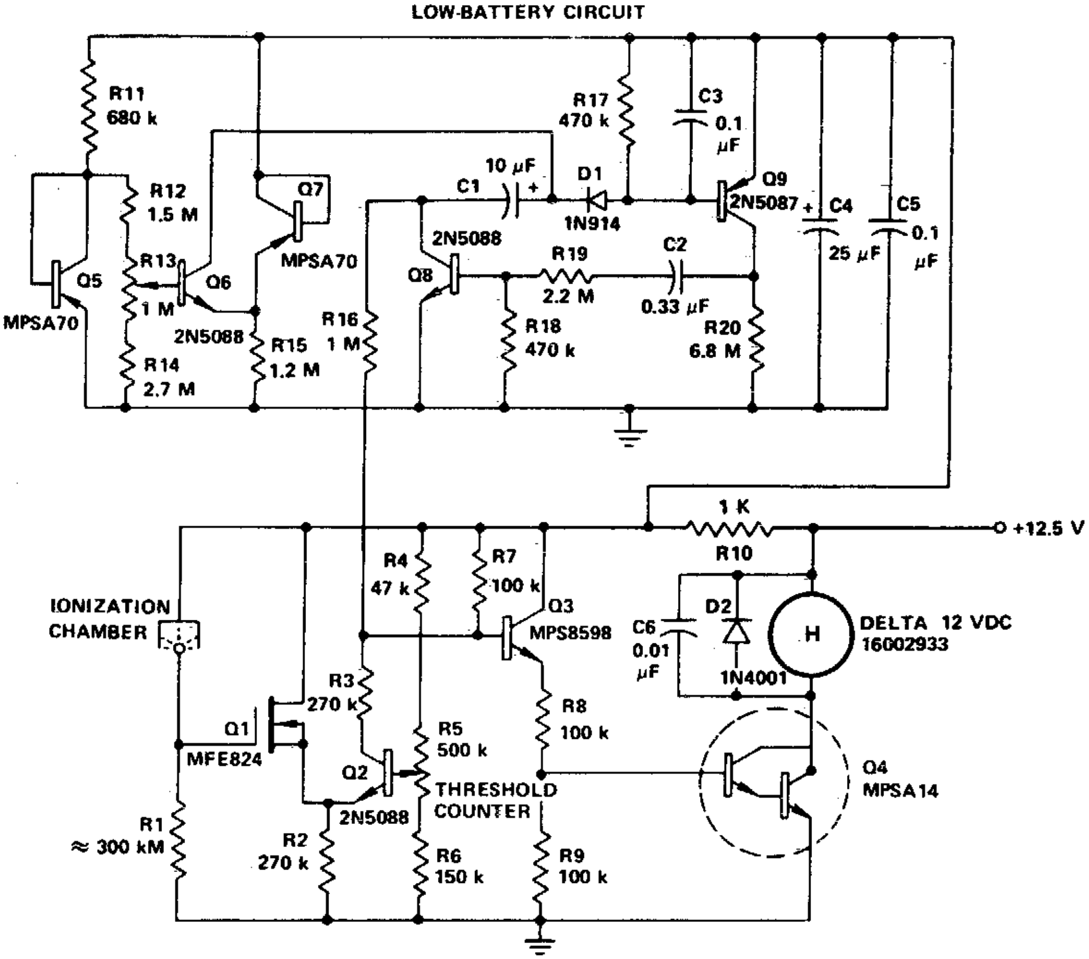

LOW-BATTERY CIRCUIT
R11
680 k
R12
1.5 M
R13
1 M
R14
2.7 M
Q5
MPSA70
Q6
2N5088
Q7
MPSA70
R15
1.2 M
R16
1 M
R17
470 k
C3
0.1 µF
10 µF
C1
D1
1N914
Q9
2N5087
C4
25 µF
C5
0.1 µF
2N5088
Q8
R19
2.2 M
C2
0.33 µF
R18
470 k
R20
6.8 M
1 K
R10
+12.5 V
IONIZATION CHAMBER
R4
47 k
R7
100 k
Q3
MPS8598
C6
0.01 µF
D2
1N4001
H
DELTA 12 VDC
16002933
R3
270 k
Q1
MFE824
R5
500 k
R8
100 k
Q2
2N5088
THRESHOLD COUNTER
Q4
MPSA14
R1
≈ 300 kM
R2
270 k
R6
150 k
R9
100 k

Fig. 40-3

Circuit Notes

If the smoke alarm signal must be a continuous one rather than pulsating, then the slightly less expensive, all discrete transistor version of the MC14572 may be used.

41

Indicators

The sources of the following circuits are contained in the Sources section beginning on page 730. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

Ten-Step Voltage-Level Indicator
Beat Frequency Indicator
Three-Step Level Indicator
Indicator and Alarm
Five-Step Voltage-Level Indicator
Visible Voltage Indicator
Voltage Level Detector
Zero Center Indicator for FM Receivers
Visual Zero-Beat Indicator

TEN-STEP VOLTAGE-LEVEL INDICATOR

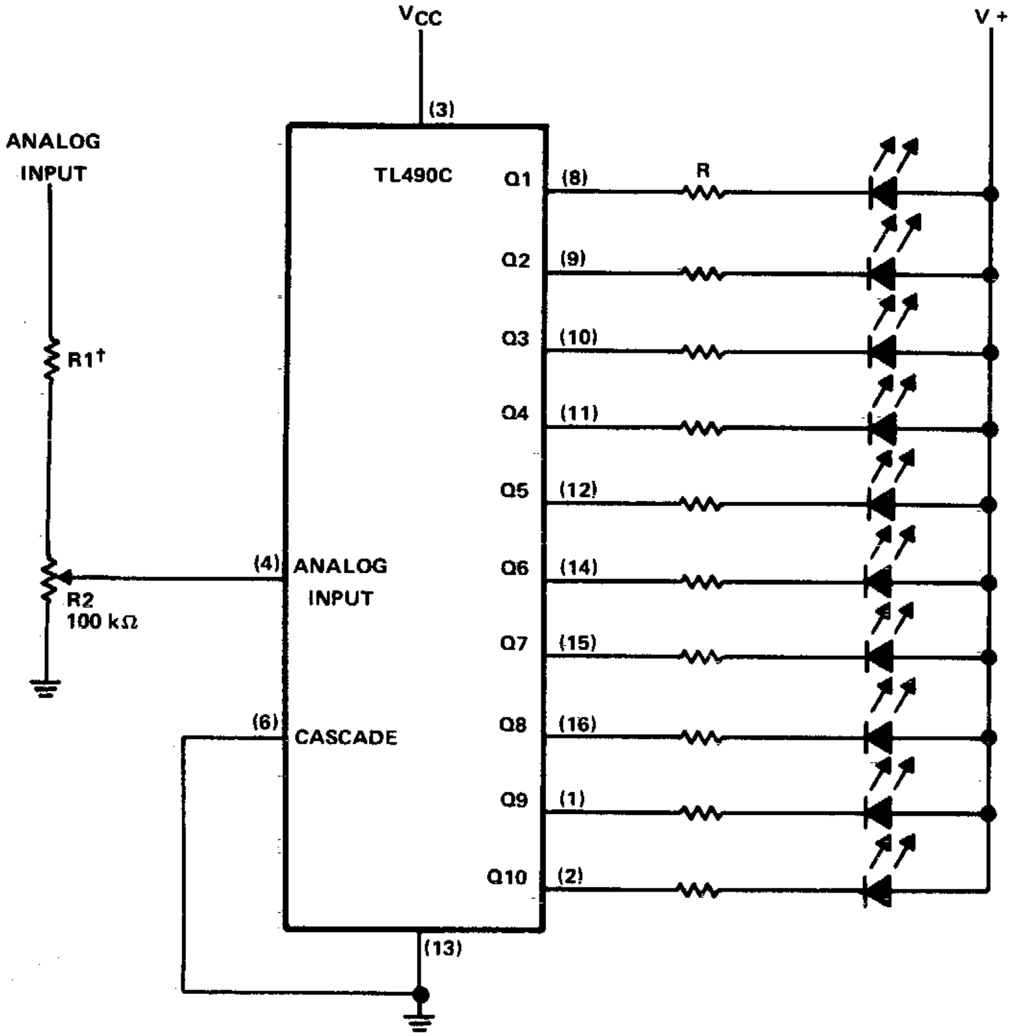

VCC
V+
(3)
ANALOG
INPUT
TL490C
Q1
(8)
R
Q2
(9)
Q3
(10)
R1†
Q4
(11)
Q5
(12)
(4)
ANALOG
INPUT
Q6
(14)
R2
100 kΩ
Q7
(15)
(6)
CASCADE
Q8
(16)
Q9
(1)
Q10
(2)
(13)

Fig. 41-1

Circuit Notes

This ten-step adjustable analog level detector is capable of sinking up to 40 milliamperes at each output. The voltage range at the input pin should range from 0 to 2 volts. Circuits of this type are useful as liquid-level indicators, pressure indicators, and temperature indicators. They may also be used with a set of active filters to provide a visual indication of harmonic content of audio signals.

BEAT FREQUENCY INDICATOR

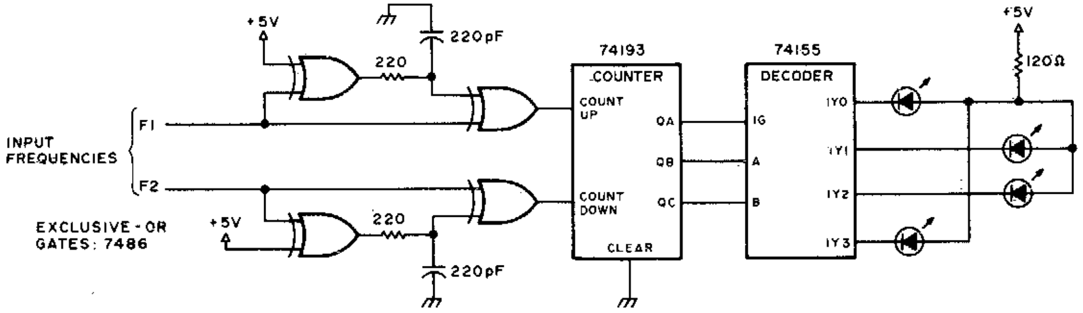

+5V
220pF
220
74193
COUNTER
COUNT UP
QA
QB
QC
COUNT DOWN
CLEAR
74155
DECODER
1G
A
B
1Y0
1Y1
1Y2
1Y3
+5V
120Ω
INPUT FREQUENCIES
F1
F2
EXCLUSIVE-OR GATES: 7486
+5V
220
220pF

Fig. 41-2

Circuit Notes

This circuit uses LEDs to display the beat frequency of two-tone oscillators. Only one LED is on at a time, and the apparent rotation of the dot is an exact indication of the best frequency. When f1 is greater than f2, a dot of light rotates clockwise; when f1 is less than f2, the dot rotates counterclockwise; and when f1 equals f2, there is no rotation.

THREE-STEP LEVEL INDICATOR

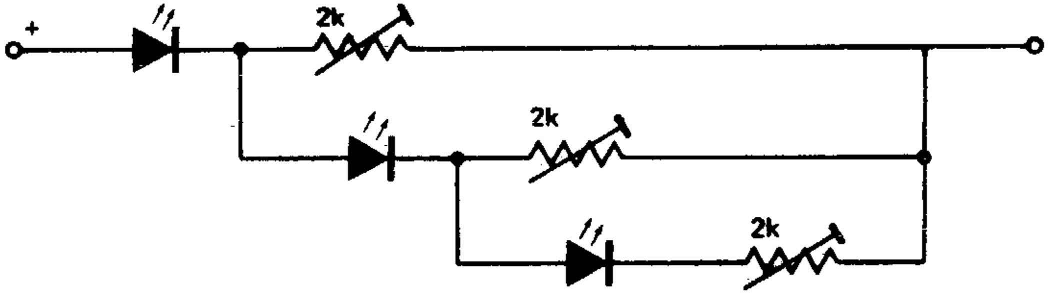
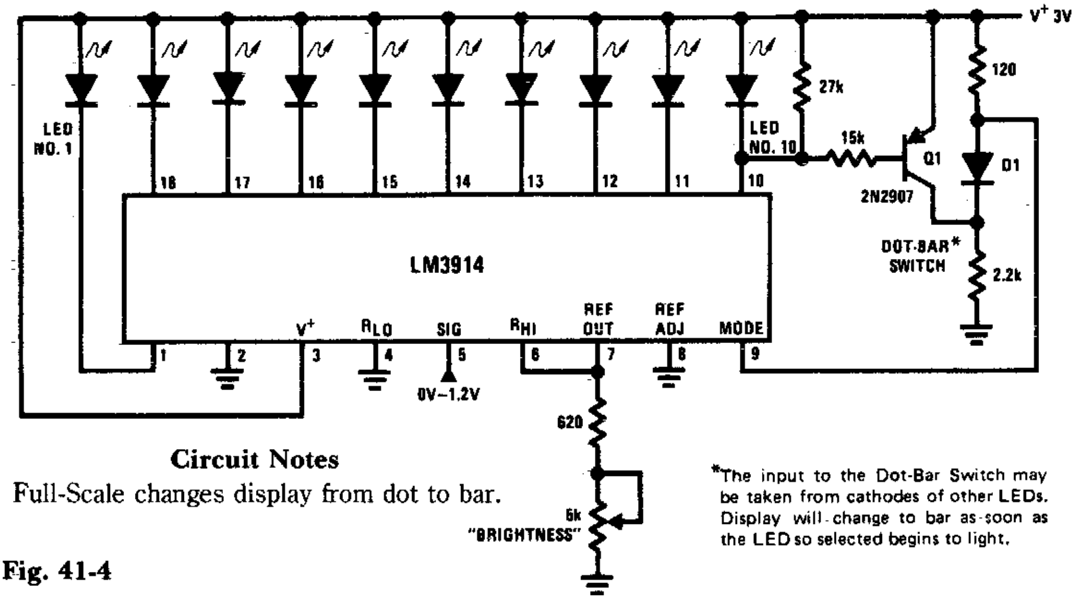


Fig. 41-3

Circuit Notes

This circuit makes a very compact level indicator where a meter would be impractical or not justified due to cost. Resistor values will depend on type of LED used. For MV50 LEDs the resistors are 2 K for steps of approx 2 V and current drain with all three LEDs on of 5 mA. The chain can be extended but current drain increases rapidly and the first LED carries all the current drawn from the supply.

INDICATOR AND ALARM



V+ 3V
LED NO. 1
18
17
16
15
14
13
12
11
10
LED NO. 10
27k
120
15k
Q1
D1
2N2907
LM3914
DOT-BAR* SWITCH
2.2k
V+
RLO
SIG
RHI
REF OUT
REF ADJ
MODE
1
2
3
4
5
6
7
8
9
0V–1.2V
620
5k
"BRIGHTNESS"

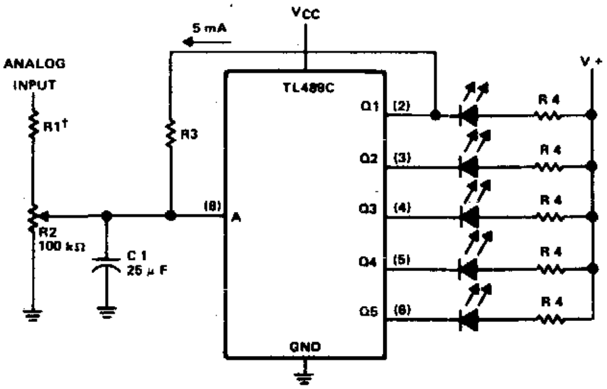
Circuit Notes

Full-Scale changes display from dot to bar.

*The input to the Dot-Bar Switch may be taken from cathodes of other LEDs. Display will change to bar as soon as the LED so selected begins to light.

Fig. 41-4

FIVE-STEP VOLTAGE-LEVEL INDICATOR



VCC
5 mA
ANALOG INPUT
R1†
R3
R2 100 kΩ
C1 25 μF
(8)
A
TL489C
Q1 (2)
Q2 (3)
Q3 (4)
Q4 (5)
Q5 (6)
GND
V+
R4
R4
R4
R4
R4

† R1 is chosen to ensure that the voltage across R2 is less than 8 volts. Normally it will be set to 1 volt.

Fig. 41-5

Circuit Notes

This circuit provides a visual indication of the input analog voltage level. It has a high input impedance at pin 8 and open-collector outputs capable of sinking up to 40 milliamperes. It is suitable for driving a linear array of 5 LEDs to indicate the level is 5 steps. The voltage at the analog input should be in the range of zero to approximately one volt and should never exceed eight volts.

VISIBLE VOLTAGE INDICATOR

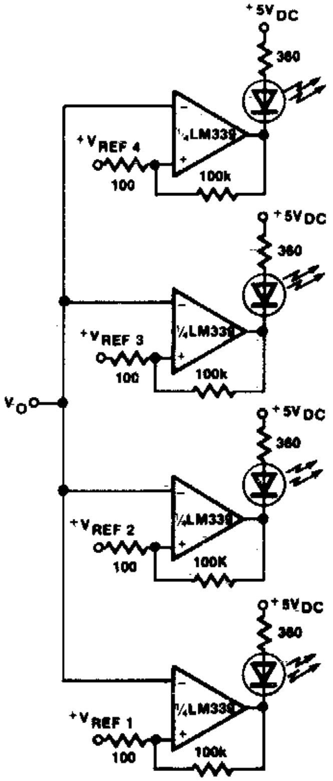

+5VDC
360
¼LM339
+VREF 4
100
100k
+5VDC
360
¼LM339
+VREF 3
100
100k
VO
+5VDC
360
¼LM339
+VREF 2
100
100K
+5VDC
360
¼LM339
+VREF 1
100
100k

Fig. 41-6

VOLTAGE LEVEL DETECTOR

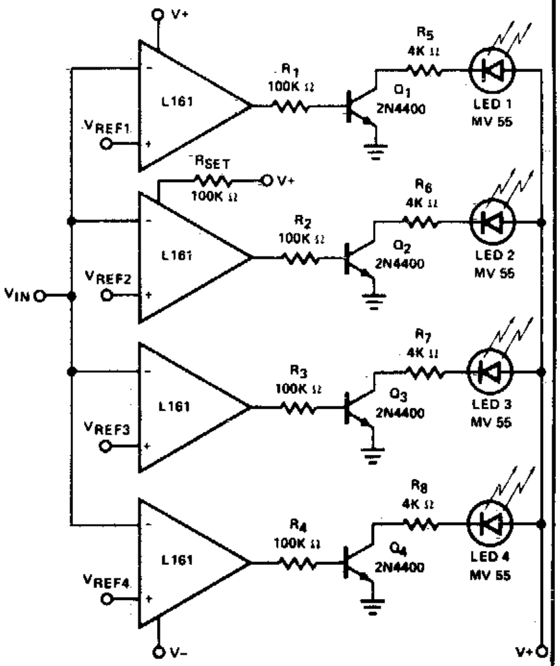

V+
R5
4K Ω
R1
100K Ω
Q1
2N4400
LED 1
MV 55
L161
VREF1
RSET
100K Ω
V+
R6
4K Ω
R2
100K Ω
Q2
2N4400
LED 2
MV 55
L161
VREF2
VIN
R7
4K Ω
R3
100K Ω
Q3
2N4400
LED 3
MV 55
L161
VREF3
R8
4K Ω
R4
100K Ω
Q4
2N4400
LED 4
MV 55
L161
VREF4
V-
V+

Fig. 41-8

ZERO CENTER INDICATOR FOR FM RECEIVERS

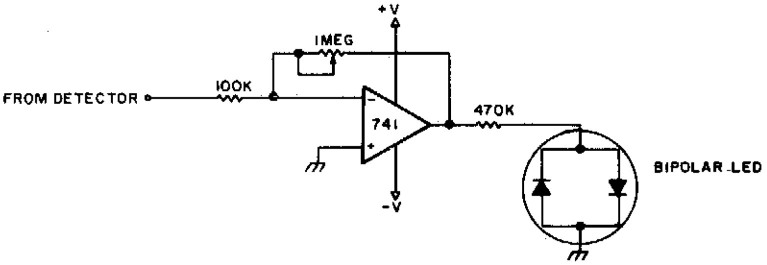
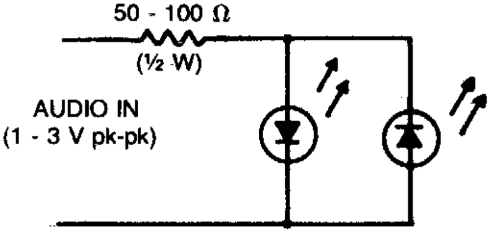

+V
IMEG
FROM DETECTOR
100K
741
470K
BIPOLAR LED
-V

Fig. 41-7

Circuit Notes

To adjust, tune in a station and adjust the 1 M pot for a null. Then ask the station to modulate and fine adjust so modulation peaks don't light the LEDs. Stations are properly tuned when neither LED is lit.

VISUAL ZERO-BEAT INDICATOR



LEDs: FAIRCHILD FLV-100 RED,
OR MONSANTO MV-5094 RED/RED,
OR MONSANTO MV-5491 RED/GREEN

Fig. 41-9

Circuit Notes

Light-emitting diodes connected with reverse polarity provide a visual indication of zero-beat frequency. Each LED is on for only half a cycle of the input. When the input frequency is more than 1 kilohertz away from the zero-beat frequency, both LEDs appear to be on all the time. As the input frequency comes within about 20 hertz of zero beat, the LEDs will flicker until zero beat is reached. Both LEDs glow or flicker until zero beat is reached, when they go out.

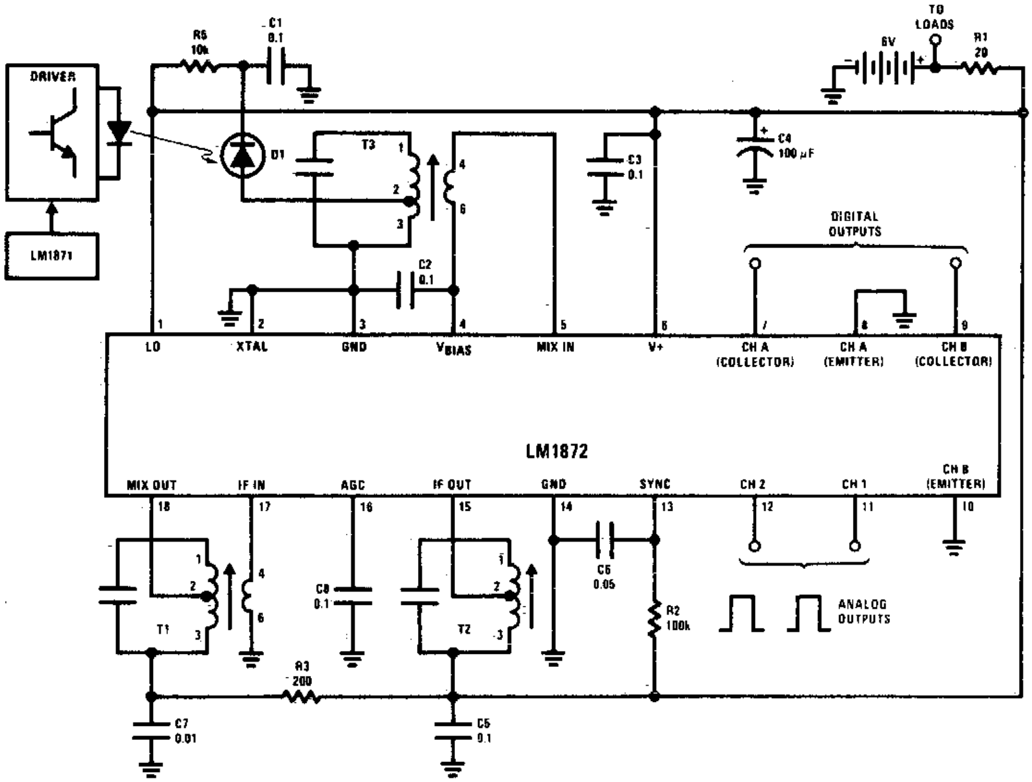
42

Infrared Circuits

The sources of the following circuits are contained in the Sources section beginning on page 730. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

IR Type Data Link
IR Remote Control Transmitter/Receiver
Compact IR Receiver
IR Transmitter
Remote Loudspeaker Via IR Link
Proximity Detector

IR TYPE DATA LINK



DRIVER
LM1871
R5
10k
C1
0.1
D1
T3
1
2
3
4
6
C2
0.1
C3
0.1
C4
100 μF
6V
TO LOADS
R1
20
DIGITAL OUTPUTS
1
2
3
4
5
6
7
8
9
LO
XTAL
GND
VBIAS
MIX IN
V+
CH A (COLLECTOR)
CH A (EMITTER)
CH B (COLLECTOR)
LM1872
MIX OUT
IF IN
AGC
IF OUT
GND
SYNC
CH 2
CH 1
CH B (EMITTER)
18
17
16
15
14
13
12
11
10
T1
T2
C8
0.1
C6
0.05
R2
100k
ANALOG OUTPUTS
R3
200
C7
0.01
C5
0.1

R1 - Load decoupling

R2 - Sync timer; $R2 = \frac{t}{0.7\ C6}$, $R2 \leq 470k$

R3 - Preamp decoupling

R5 - Photodiode decoupling

C1 - Photodiode decoupling

C2 - V_{BIAS} bypass

C3 - V^{+} bypass

C4 - Load decoupling

C5 - IF bypass; optional

C6 - Sync timer; $C6 = \frac{t_{SYNC}}{0.7\ R2}$, $C6 \leq 0.5\ \mu F$

C7 - Preamp decoupling

C8 - AGC

T1 - 455 kHz preamp transformer
Toko* 10 EZC type (RMC-502182), Qu = 110
Pin 1-2, 82T; pin 2-3, 82T
Pin 1-3, 164T; pin 4-6, 30T

T2 - 455 kHz IF transformer
Toko* 10 EZC type (RMC-402503), Qu = 110
Pin 1-2, 98T; pin 2-3, 66T
Pin 1-3, 164T; pin 4-6, 8T

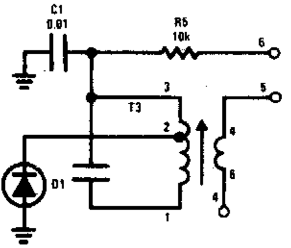
T3 - 455 kHz input transformer
Toko* 10 EZC type (RMC-202313), Qu = 110
Pin 1-2, 131T; pin 2-3, 33T
Pin 1-3, 164T; pin 4-6, 5T

D1 - PN or PIN Silicon Photodiode

BOTTOM VIEW

Photodiode, D1		Active Area (cm²)
Vactec	VTS 5088	0.18
Vactec	VTS 6089	0.52
UDT	PIN 5D or 6 DP	0.20
UDT	PIN 220 DP	2.0
Siemens	BPY 12	0.20

* Toko America, Inc.
5520 West Touhy Ave.
Skokie, Ill. 60077
(312)677-3640 Tlx: 72-4372

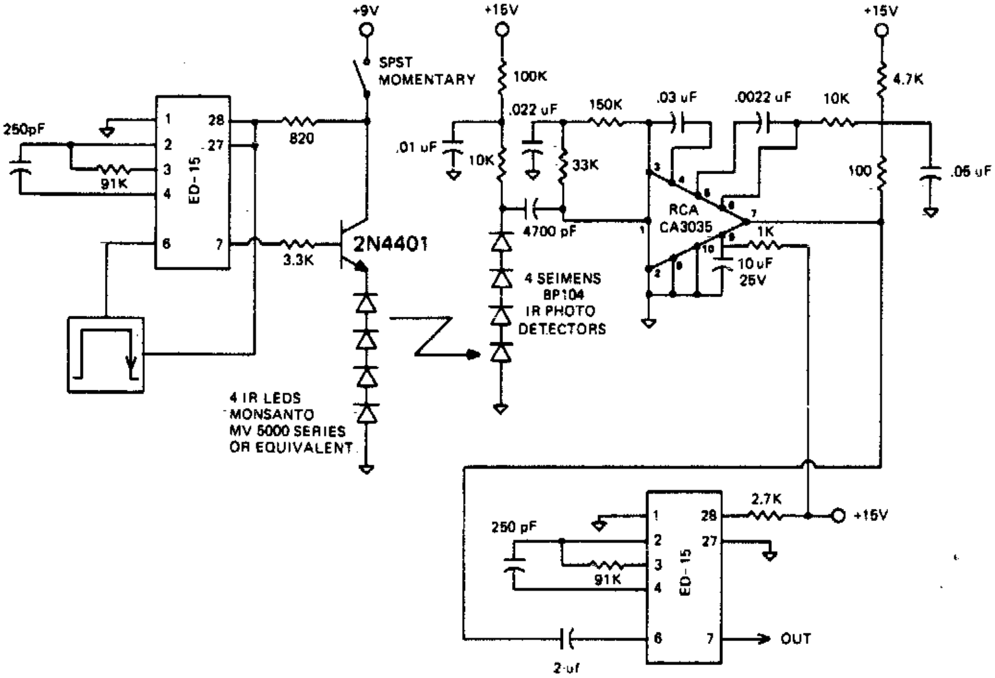


C1
0.01
R5
10k
6
5
3
T3
2
4
6
4
1
D1

Input Stage Where the Case of D1 Is Connected to the Anode

Fig. 42-1

IR REMOTE CONTROL TRANSMITTER/RECEIVER



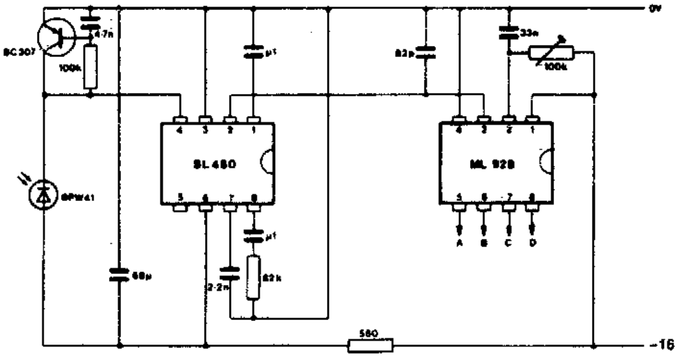
+9V
+15V
+15V
SPST MOMENTARY
100K
4.7K
250pF
ED-15
1
2
3
4
6
28
27
7
91K
820
3.3K
2N4401
.01 uF
10K
.022 uF
150K
.03 uF
.0022 uF
10K
33K
100
.05 uF
4700 pF
RCA CA3035
1K
10 uF 25V
4 SEIMENS BP104 IR PHOTO DETECTORS
4 IR LEDS MONSANTO MV 5000 SERIES OR EQUIVALENT
250 pF
91K
2.7K
+15V
OUT
2 uf

Fig. 42-2

Circuit Notes

The circuit is designed to operate at 25 kHz. The data stream turns the 2N4401 hard on or off depending upon the coded state. This in turn switches the series infrared LEDs on and off. The receiver circuit consists of a three stage amplifier with photo diodes arrayed for maximum coverage of the reception area. The range of this set-up should be about 10 meters.

COMPACT IR RECEIVER



0V
BC307
4.7n
100k
μ1
82p
33n
100k
SL480
ML 928
BPW41
68p
2·2n
82k
A
B
C
D
560
–15

Fig. 42-3

IR TRANSMITTER

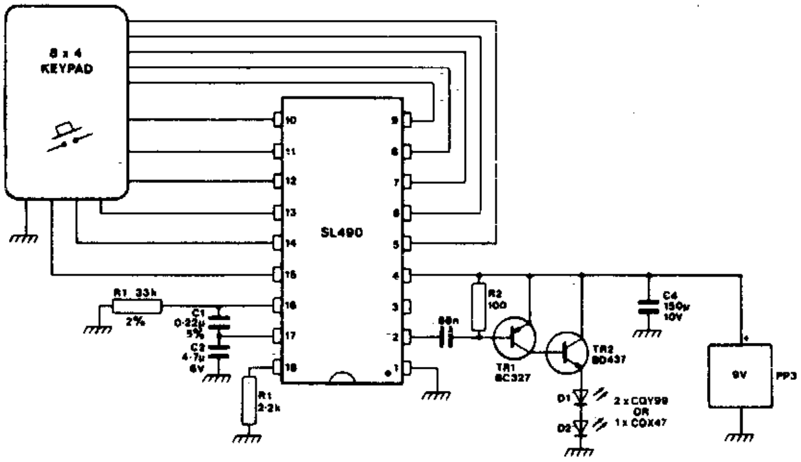

8 x 4
KEYPAD
SL490
10
11
12
13
14
15
16
17
18
9
8
7
6
5
4
3
2
1
R1 33k
2%
C1
0·22µ
5%
C2
4·7µ
6V
R1
2·2k
68n
R2
100
TR1
BC327
TR2
BD437
C4
150µ
10V
D1
D2
2 x CQY99
OR
1 x CQX47
9V
PP3
+

Fig. 42-4

Circuit Notes

This simple infra-red transmitter, where the PPM output from pin 2 of the SL490 is fed to the base of the PNP trasmitter TR1, produces an amplified current pulse about 15 µsec wide. This pulse is further amplified by TR2 and applied to the infra-red diodes D1 and D2.

REMOTE LOUDSPEAKER VIA IR LINK

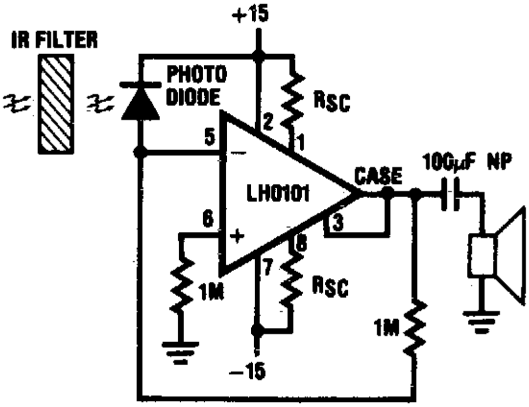

+15
IR FILTER
PHOTO
DIODE
RSC
2
1
5
−
LH0101
CASE
100µF NP
6
+
3
8
7
1M
RSC
1M
−15

Fig. 42-5

PROXIMITY DETECTOR

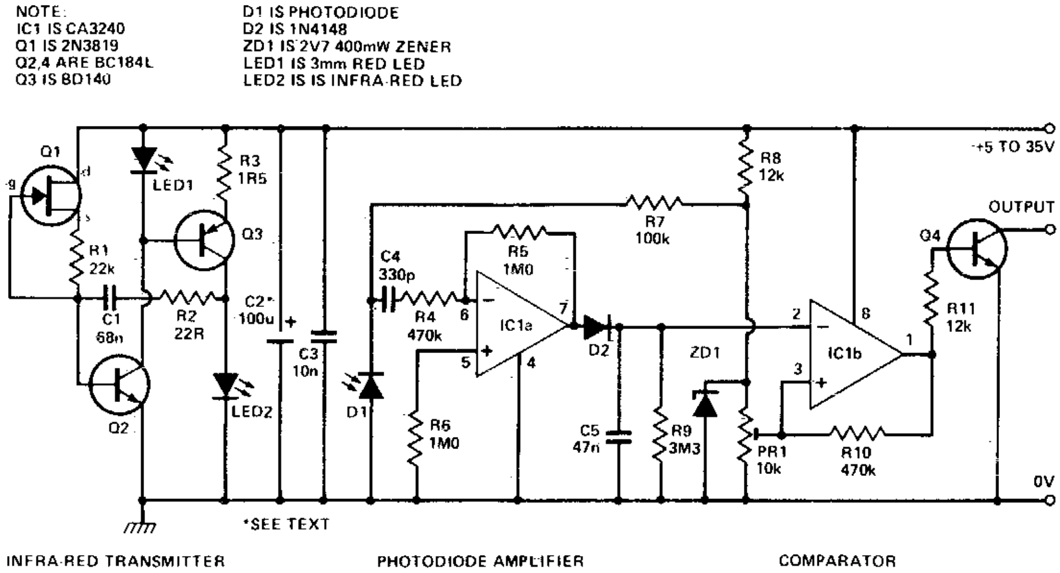

NOTE:
IC1 IS CA3240
Q1 IS 2N3819
Q2,4 ARE BC184L
Q3 IS BD140
D1 IS PHOTODIODE
D2 IS 1N4148
ZD1 IS 2V7 400mW ZENER
LED1 IS 3mm RED LED
LED2 IS IS INFRA-RED LED
Q1
g
d
s
LED1
R3
1R5
Q3
R1
22k
C1
68n
R2
22R
C2*
100u
+
C3
10n
LED2
Q2
C4
330p
R4
470k
R5
1M0
6
5
7
4
IC1a
D1
R6
1M0
D2
C5
47n
R9
3M3
ZD1
R7
100k
R8
12k
PR1
10k
2
3
8
1
IC1b
R10
470k
R11
12k
Q4
OUTPUT
+5 TO 35V
0V
*SEE TEXT
INFRA-RED TRANSMITTER
PHOTODIODE AMPLIFIER
COMPARATOR

Fig. 42-6

Circuit Notes

This circuit provides a means of detecting the presence of anything by the reflection of infra-red light and provides a direct digital output of object detection. By the use of modulation and high power bursts of infra-red at a very low duty cycle, a detection range of over a foot is achieved. Works on the principle of transmitting a beam of modulated infra-red light from the emitter diode LED2, and receiving reflections from objects passing in front of the beam with a photodiode detector D1. The circuit consists of an infra-red transmitter, photodiode amplifier, and a variable threshold comparator.

43

Instrumentation Amplifiers

The sources of the following circuits are contained in the Sources section beginning on page 730. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

Instrumentation Amplifier
Triple Op-Amp Instrumentation Amplifier
Differential Input Instrumentation Amplifier with High CMRR
Instrumentation Amplifier with High CMRR
Level-Shifting Isolation Amplifier
Variable Gain, Differential-Input Instrumentation Amplifier
Instrumentation Amplifier
Low Signal Level, High Impedance Instrumentation Amplifier
Chopper Channel Amplifier
Battery Powered Buffer Amplifier for Standard Cell
Bridge Transducer Amplifier
Instrumentation Amplifier
Isolation Amplifier for Medical Telemetry
High Gain Differential Instrumentation Amplifier
High Impedance Bridge Amplifier
Instrumentation Amplifier (Two Op Amp Design)
Instrumentation Amplifier
Differential Input Instrumentation Amplifier
High Impedance Differential Amplifier
High Speed Instrumentation Amplifier
Very High Impedance Instrumentation Amplifier
Precision FET Input Instrumentation Amplifier
High Stability Thermocouple Amplifier
High Stability Thermocouple Amplifier
High Impedance, Low Drift Instrumentation Amplifier

INSTRUMENTATION AMPLIFIER

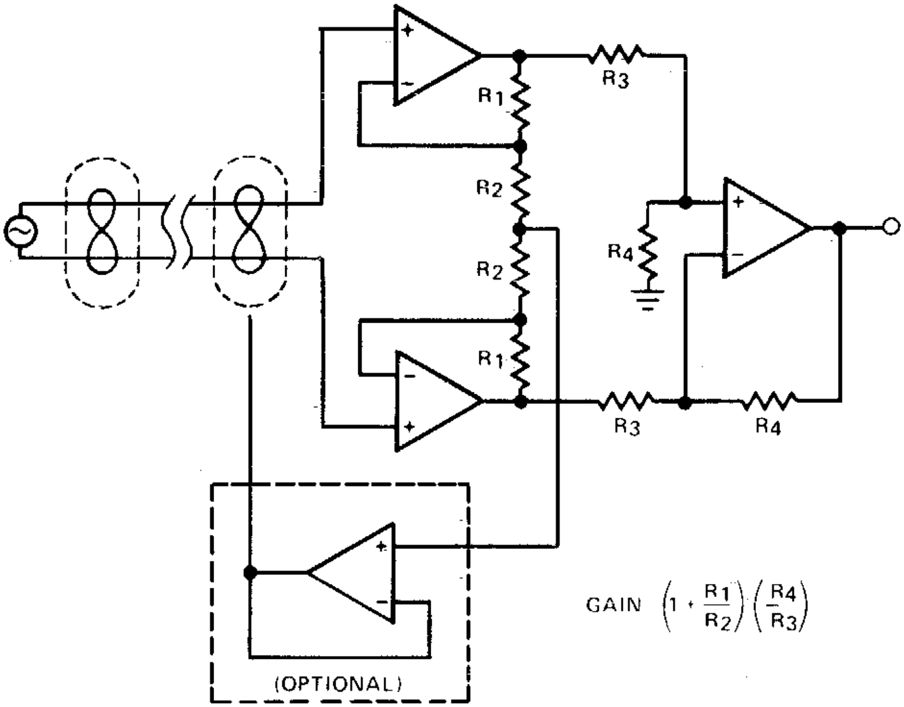


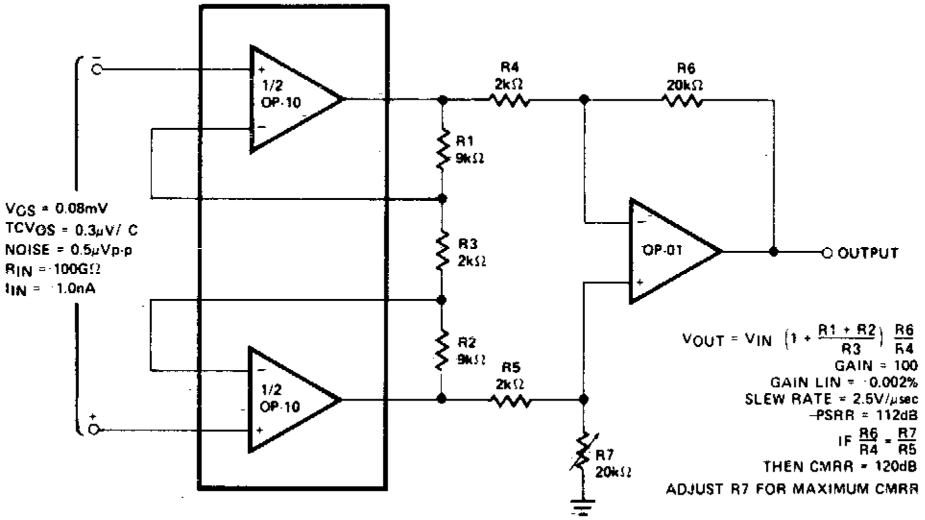
Fig. 43-1

$$\text{GAIN } \left(1 + \frac{R_1}{R_2}\right)\left(\frac{R_4}{R_3}\right)$$

Circuit Notes

Instrumentation amplifiers (differential amplifiers) are specifically designed to extract and amplify small differential signals from much larger common mode voltages. To serve as building blocks in instrumentation amplifiers, op amps must have very low offset voltage drift, high gain and wide bandwidth. The HA-4620/5604 is suited for this application. The optional circuitry makes use of the fourth amplifier section as a shield driver which enhances the ac common mode rejection by nullifying the effects of capacitance-to-ground mismatch between input conductors.

TRIPLE OP-AMP INSTRUMENTATION AMPLIFIER



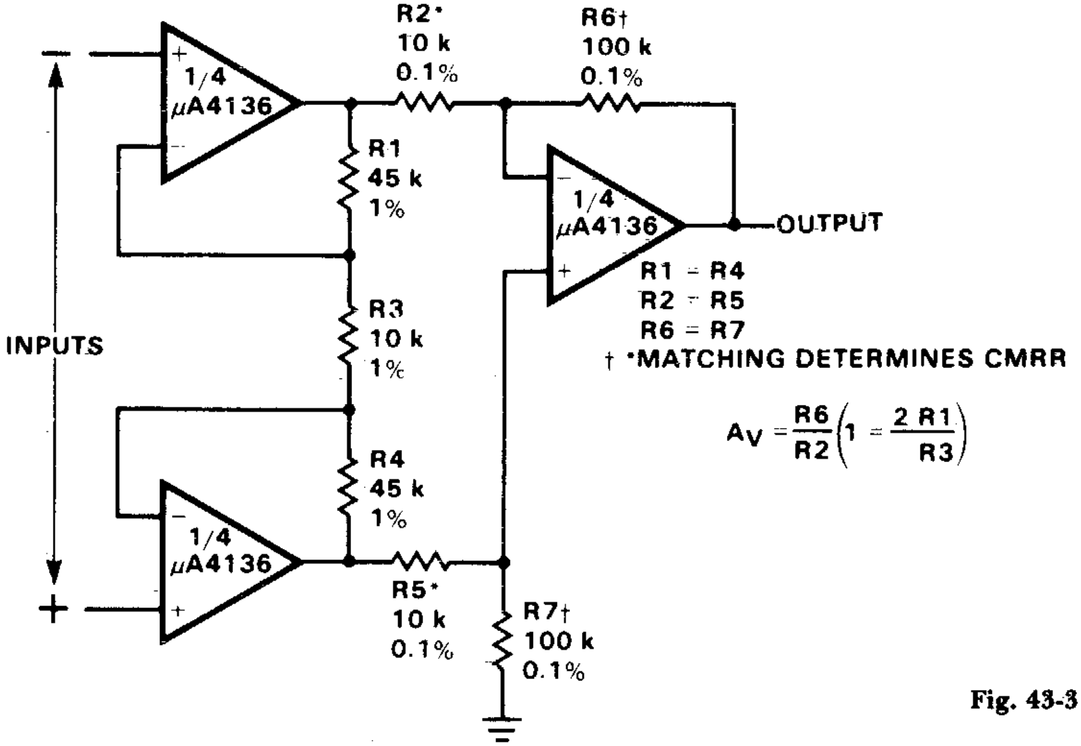
1/2 OP-10
1/2 OP-10
OP-01
R4 2kΩ
R6 20kΩ
R1 9kΩ
R3 2kΩ
R2 9kΩ
R5 2kΩ
R7 20kΩ
OUTPUT
VOS = 0.08mV
TCVOS = 0.3μV/°C
NOISE = 0.5μVp-p
RIN = 100GΩ
IIN = 1.0nA
GAIN = 100
GAIN LIN = 0.002%
SLEW RATE = 2.5V/μsec
PSRR = 112dB
THEN CMRR = 120dB
ADJUST R7 FOR MAXIMUM CMRR

$$V_{OUT} = V_{IN}\left(1 + \frac{R1 + R2}{R3}\right)\frac{R6}{R4}$$

$$\text{IF } \frac{R6}{R4} = \frac{R7}{R5}$$

Fig. 43-2

DIFFERENTIAL INPUT INSTRUMENTATION AMPLIFIER WITH HIGH COMMON MODE REJECTION

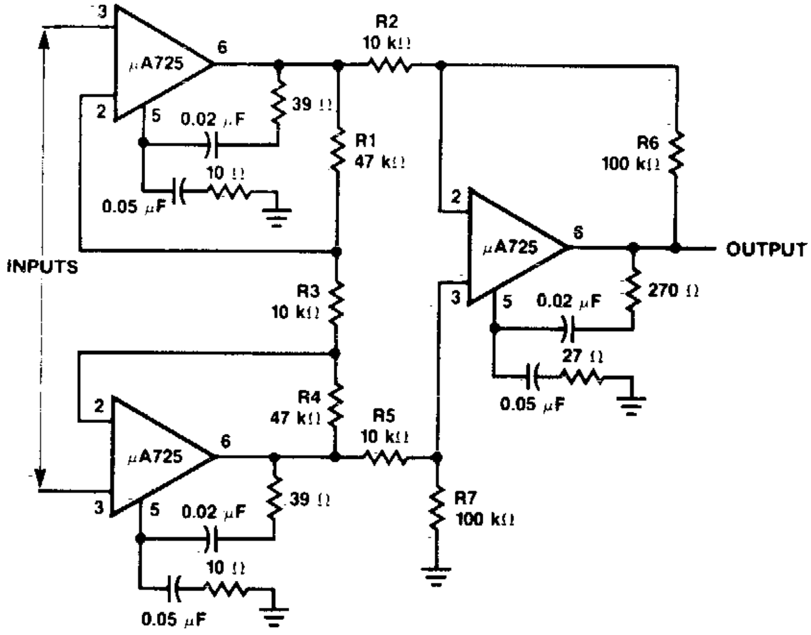


1/4 μA4136
1/4 μA4136
1/4 μA4136
R2* 10 k 0.1%
R6† 100 k 0.1%
R1 45 k 1%
R3 10 k 1%
R4 45 k 1%
R5* 10 k 0.1%
R7† 100 k 0.1%
INPUTS
OUTPUT
R1 = R4
R2 = R5
R6 = R7
† *MATCHING DETERMINES CMRR

$$A_V = \frac{R6}{R2}\left(1 = \frac{2\,R1}{R3}\right)$$

Fig. 43-3

INSTRUMENTATION AMPLIFIER WITH HIGH COMMON MODE REJECTION



3
μA725
6
2
5
0.02 μF
39 Ω
10 Ω
0.05 μF
R2
10 kΩ
R1
47 kΩ
INPUTS
R3
10 kΩ
R4
47 kΩ
R5
10 kΩ
2
μA725
6
3
5
0.02 μF
39 Ω
10 Ω
0.05 μF
R7
100 kΩ
R6
100 kΩ
2
μA725
6
3
5
0.02 μF
270 Ω
27 Ω
0.05 μF
OUTPUT

$\frac{R1}{R6} = \frac{R3}{R4}$ for best CMRR

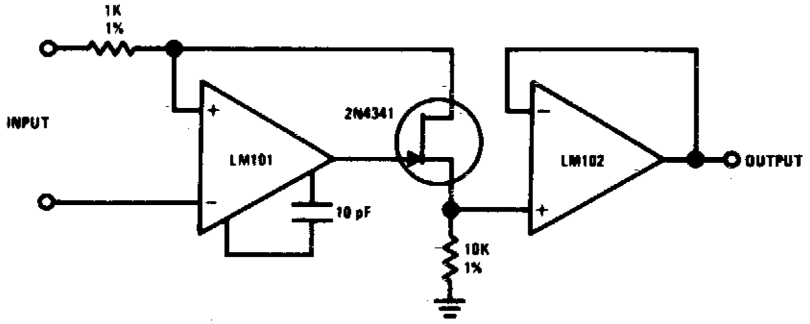
$R3 = R4$

$R1 = R6 = 10\,R3$

$\text{Gain} = \frac{R6}{R7}$

Fig. 43-4

LEVEL-SHIFTING ISOLATION AMPLIFIER



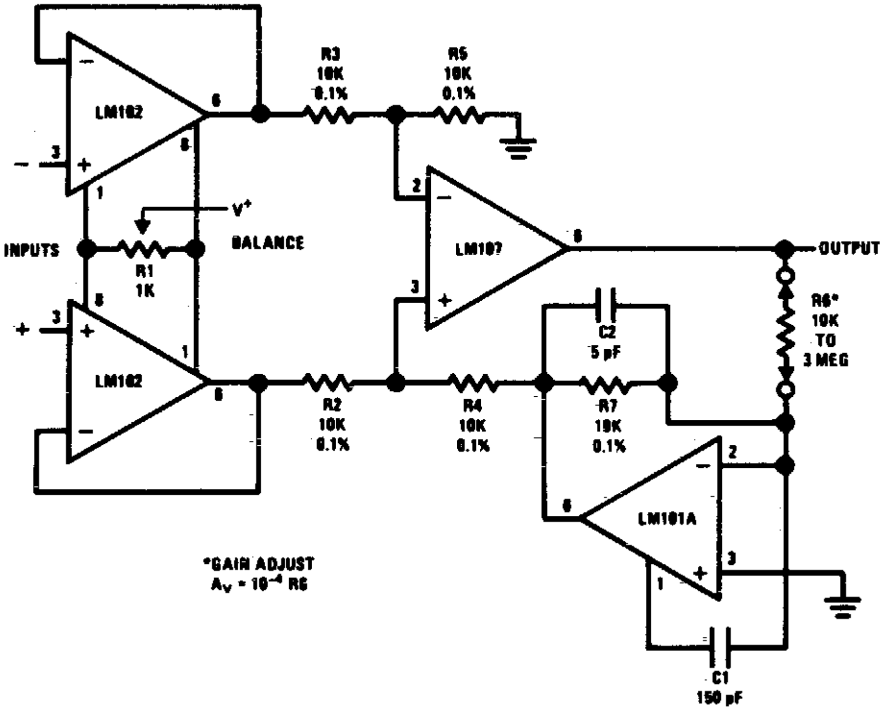
1K
1%
INPUT
+
−
LM101
10 pF
2N4341
10K
1%
−
+
LM102
OUTPUT

Fig. 43-5

Circuit Notes

The 2N4341 JFET is used as a level shifter between two op amps operated at different power supply voltages. The JFET is ideally suited for this type of application because $I_D = I_S$.

VARIABLE GAIN, DIFFERENTIAL-INPUT INSTRUMENTATION AMPLIFIER


LM102
LM102
LM107
LM101A
INPUTS
−
+
3
6
8
1
2
R1
1K
V+
BALANCE
R3
10K
0.1%
R5
10K
0.1%
R2
10K
0.1%
R4
10K
0.1%
R7
10K
0.1%
C2
5 pF
R6*
10K
TO
3 MEG
OUTPUT
C1
150 pF

*GAIN ADJUST
$A_V = 10^{-4}$ R6

Fig. 43-6

INSTRUMENTATION AMPLIFIER

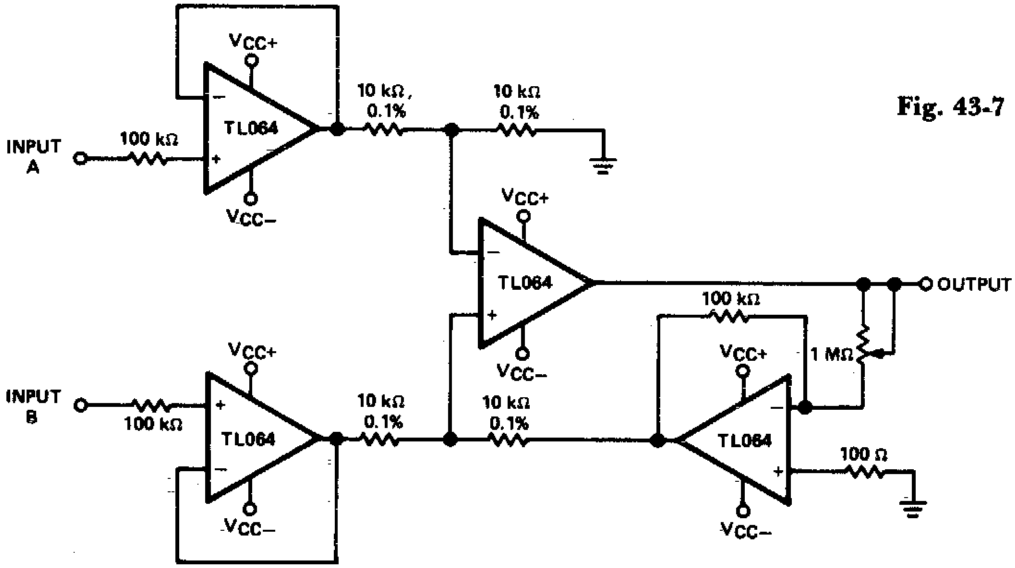

INPUT A
INPUT B
100 kΩ
100 kΩ
TL064
TL064
TL064
TL064
VCC+
VCC−
10 kΩ, 0.1%
10 kΩ 0.1%
10 kΩ 0.1%
10 kΩ 0.1%
100 kΩ
1 MΩ
100 Ω
OUTPUT

Fig. 43-7

LOW SIGNAL LEVEL, HIGH IMPEDANCE INSTRUMENTATION AMPLIFIER

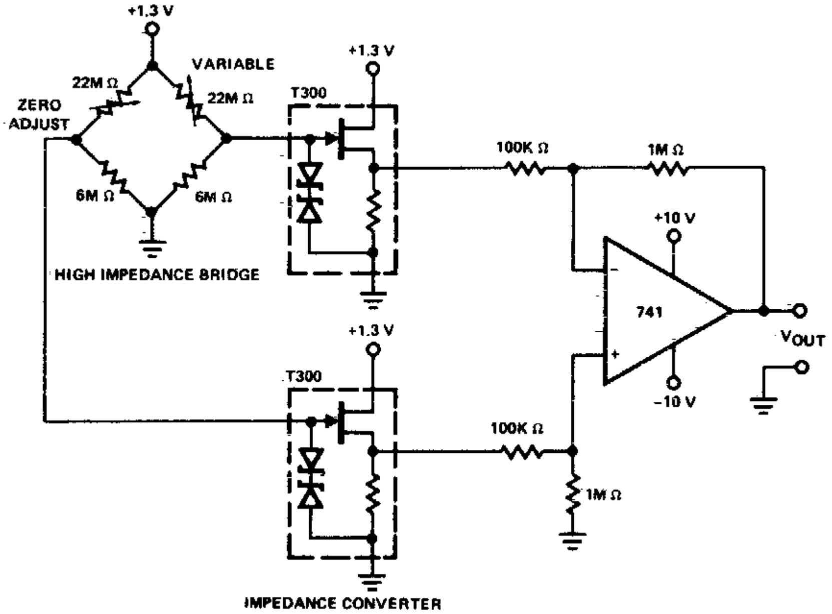

+1.3 V
VARIABLE
22M Ω
22M Ω
ZERO ADJUST
6M Ω
6M Ω
HIGH IMPEDANCE BRIDGE
T300
+1.3 V
100K Ω
1M Ω
+10 V
741
−10 V
VOUT
+1.3 V
T300
100K Ω
1M Ω
IMPEDANCE CONVERTER

Fig. 43-8

CHOPPER CHANNEL AMPLIFIER

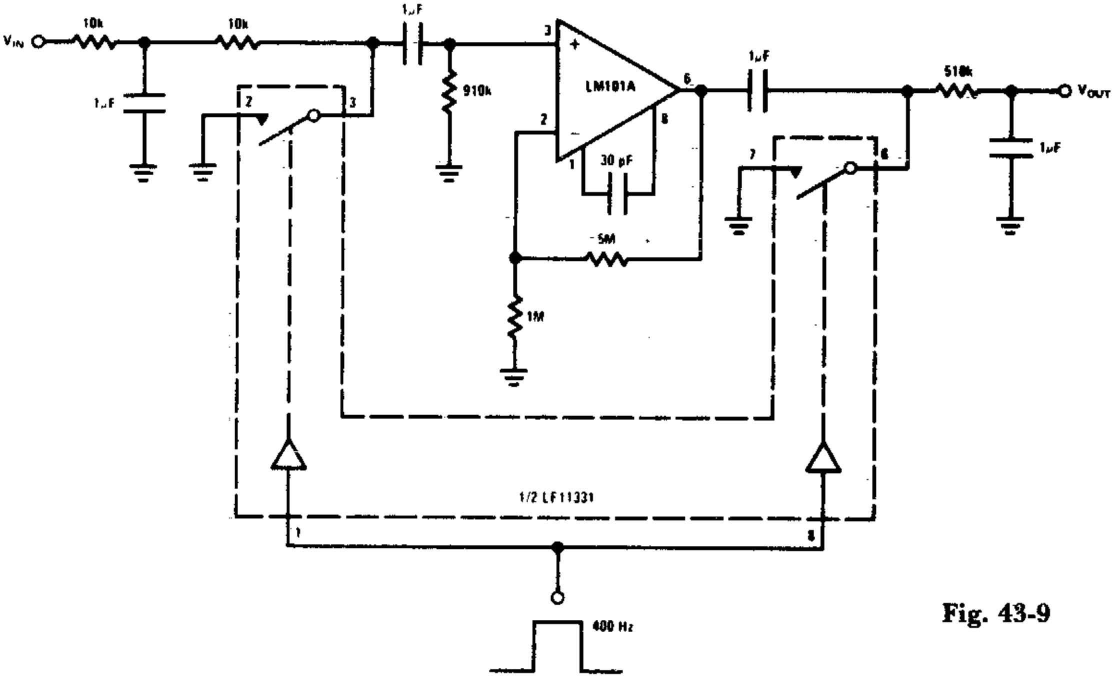
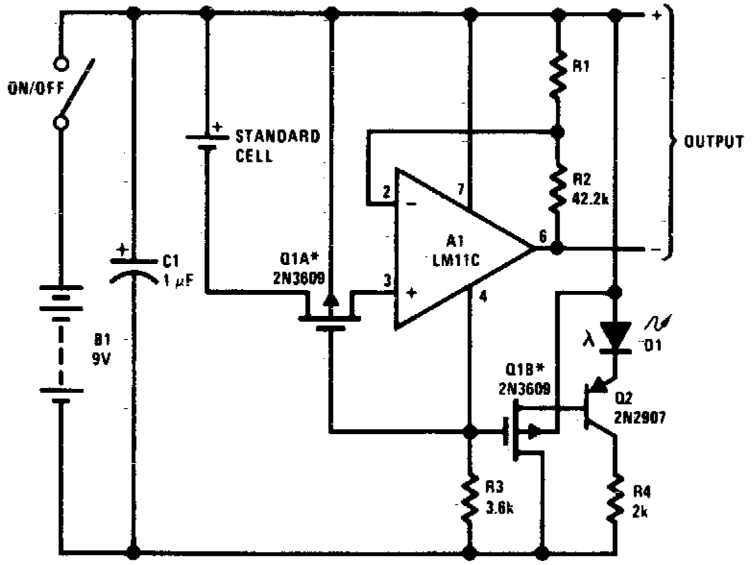

VIN
10k
10k
1μF
1μF
910k
LM101A
30 pF
5M
1M
1μF
510k
1μF
VOUT
1/2 LF11331
400 Hz

Fig. 43-9

BATTERY POWERED BUFFER AMPLIFIER FOR STANDARD CELL


ON/OFF
B1
9V
C1
1 μF
STANDARD
CELL
Q1A*
2N3609
A1
LM11C
2
3
4
6
7
R1
R2
42.2k
OUTPUT
+
−
D1
λ
Q1B*
2N3609
Q2
2N2907
R3
3.6k
R4
2k

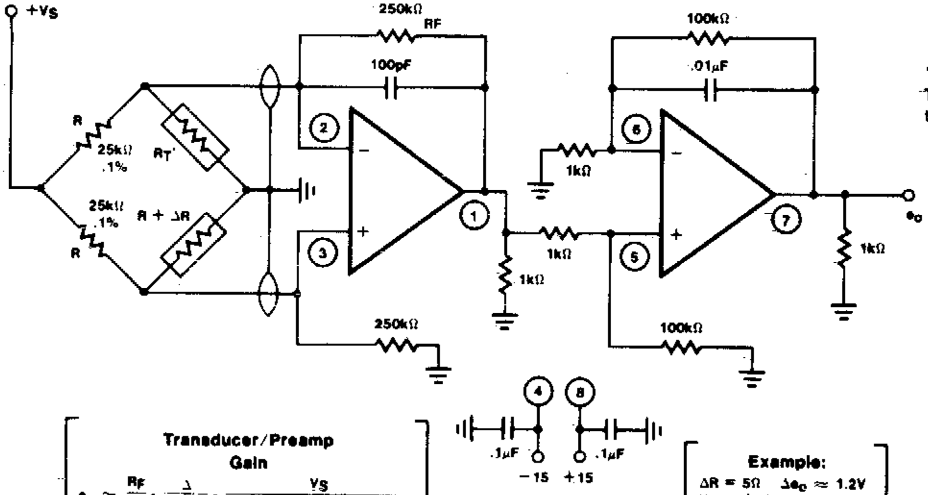
* cannot have gate protection diode; $V_{TH} > V_{OUT}$

Fig. 43-10

Circuit Notes

This circuit has negligible loading and disconnects the cell for low supply voltage or overload on output. The indicator diode extinguishes as disconnect circuitry is activated.

BRIDGE TRANSDUCER AMPLIFIER


+Vs
R
25kΩ
.1%
RT'
25kΩ
.1%
R + ΔR
R
250kΩ
RF
100pF
②
③
①
1kΩ
1kΩ
250kΩ
100kΩ
.01μF
⑥
⑤
⑦
1kΩ
100kΩ
1kΩ
eo
④
⑧
.1μF
.1μF
−15
+15

*NOTE
Thermal compensation transducer (non-active)

Transducer/Preamp Gain

$$e_o \cong \frac{R_F}{R} \cdot \frac{\Delta}{1 + \Delta} \cdot \frac{V_S}{(2 + \Delta)/[(1 + \Delta) + R/R_F]}$$

$$\Delta = \frac{\Delta R}{R}$$

IC = NE/SE5512

Example:

$\Delta R = 5\Omega \quad \Delta e_o \approx 1.2V$
$V_S = +10V$

Fig. 43-11

INSTRUMENTATION AMPLIFIER

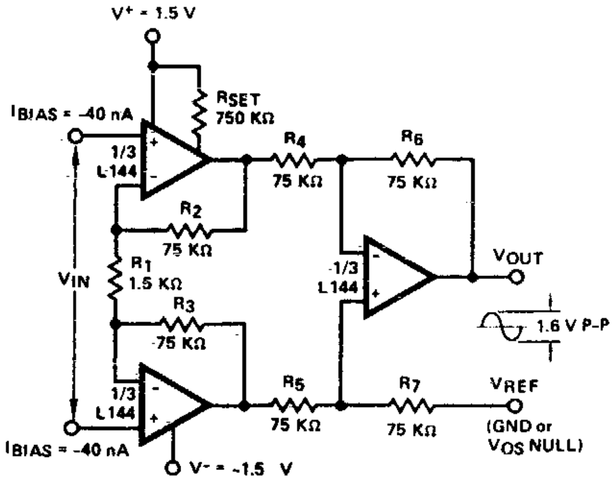

V+ = 1.5 V
RSET
750 KΩ
IBIAS = -40 nA
1/3
L144
R4
75 KΩ
R6
75 KΩ
R2
75 KΩ
R1
1.5 KΩ
VIN
R3
75 KΩ
1/3
L144
VOUT
1.6 V P-P
1/3
L144
R5
75 KΩ
R7
75 KΩ
VREF
(GND or
VOS NULL)
IBIAS = -40 nA
V- = -1.5 V

Fig. 43-12

$A_V = 1 + \frac{2R_2}{R_1}$

$P_D = 135\ \mu W$

V_{OS} (TYP) RTI = 0.45 mV

Circuit Notes

Three-amplifier circuit consumes only 135 μW of power from a ±1.5 V power supply. With a gain of 101, the instrumentation amplifier is ideal in sensor interface and biomedical preamplifier applications. The first stage provides all of the gain while the second stage is used to provide common mode rejection and double-ended to single-ended conversion.

ISOLATION AMPLIFIER FOR MEDICAL TELEMETRY

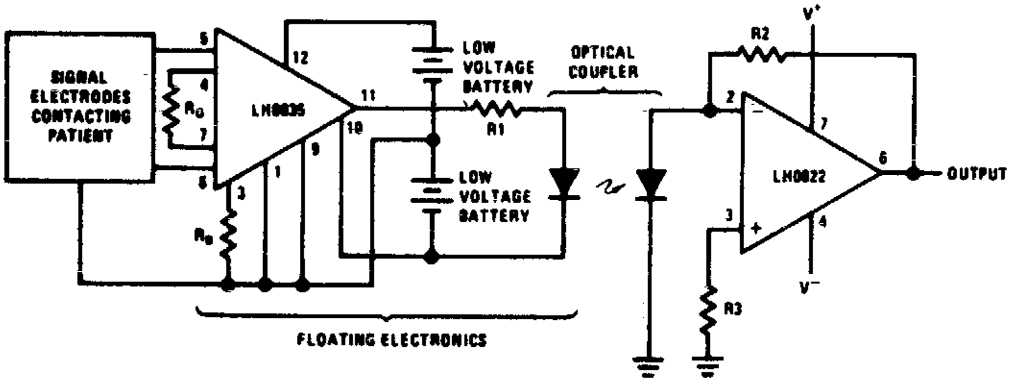

SIGNAL
ELECTRODES
CONTACTING
PATIENT
5
4
RG
7
6
LH0036
12
11
10
9
1
3
RS
LOW
VOLTAGE
BATTERY
R1
LOW
VOLTAGE
BATTERY
OPTICAL
COUPLER
R2
V+
2
7
LH0022
6
OUTPUT
3
4
V−
R3
FLOATING ELECTRONICS

Fig. 43-13

HIGH GAIN DIFFERENTIAL INSTRUMENTATION AMPLIFIER

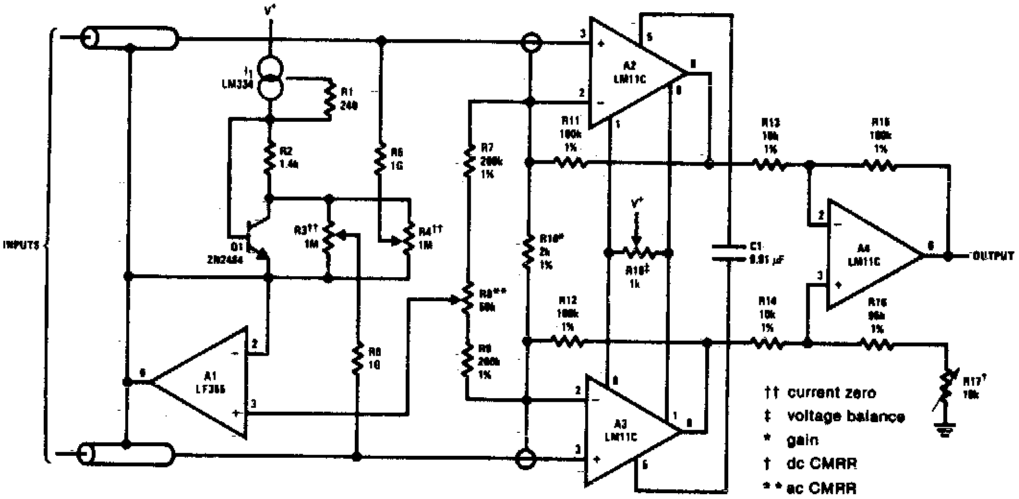

V+
I1
LM334
R1
240
R2
1.4k
R6
1G
R7
200k
1%
R11
100k
1%
R13
10k
1%
R15
100k
1%
INPUTS
Q1
2N2484
R3††
1M
R4††
1M
R10*
2k
1%
R18‡
1k
C1
0.01 μF
A2
LM11C
A3
LM11C
A4
LM11C
A1
LF355
OUTPUT
R8
1G
R8**
50k
R9
200k
1%
R12
100k
1%
R14
10k
1%
R16
95k
1%
R17†
10k
†† current zero
‡ voltage balance
* gain
† dc CMRR
** ac CMRR

Fig. 43-14

Circuit Notes

This circuit includes input guarding, cable bootstrapping, and bias current compensation. Differential bandwidth is reduced by C1 which also makes common-mode rejection less dependent on matching of input amplifiers.

HIGH IMPEDANCE BRIDGE AMPLIFIER

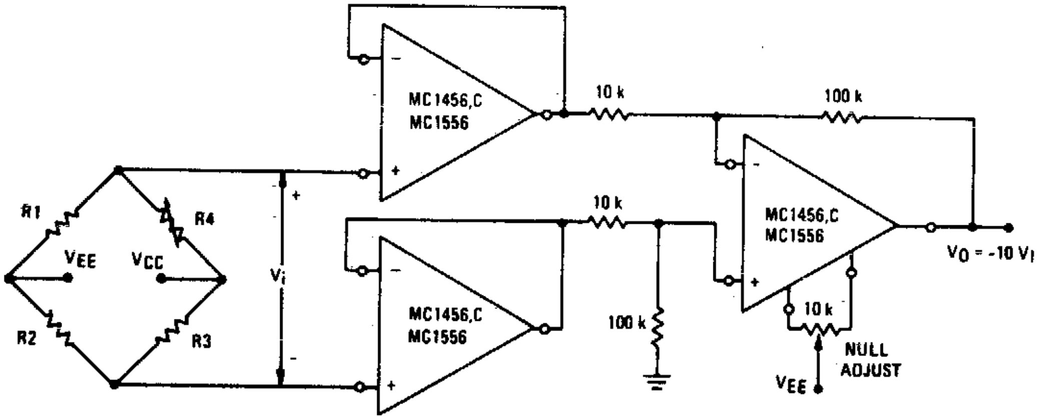
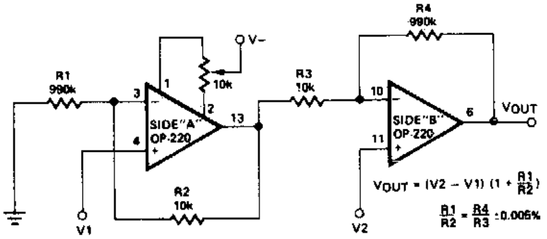

MC1456,C
MC1556
10 k
100 k
R1
R4
VEE
VCC
Vi
+
-
R2
R3
10 k
MC1456,C
MC1556
Vo = -10 Vi
MC1456,C
MC1556
100 k
10 k
NULL
ADJUST
VEE

Fig. 43-15

INSTRUMENTATION AMPLIFIER (TWO OP AMP DESIGN)


R1
990k
3
1
10k
V−
2
SIDE "A"
OP-220
4
13
R3
10k
R4
990k
10
SIDE "B"
OP-220
11
6
VOUT
R2
10k
V1
V2

$V_{OUT} = (V2 - V1)\left(1 + \frac{R1}{R2}\right)$

$\frac{R1}{R2} = \frac{R4}{R3} \pm 0.005\%$

Fig. 43-16

INSTRUMENTATION AMPLIFIER

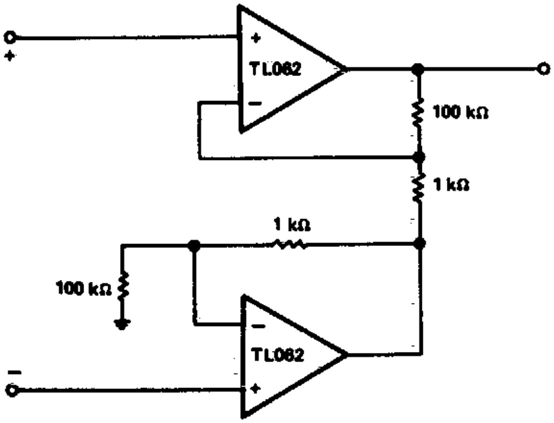
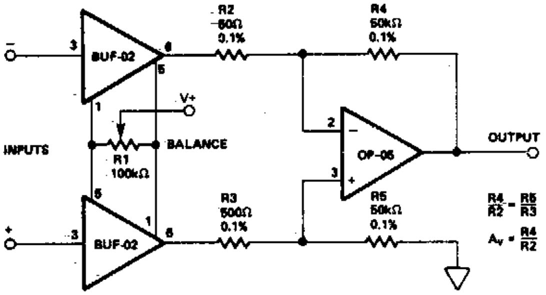

+
TL062
−
100 kΩ
1 kΩ
1 kΩ
100 kΩ
−
TL062
+
−
+

Fig. 43-17

DIFFERENTIAL INPUT INSTRUMENTATION AMPLIFIER

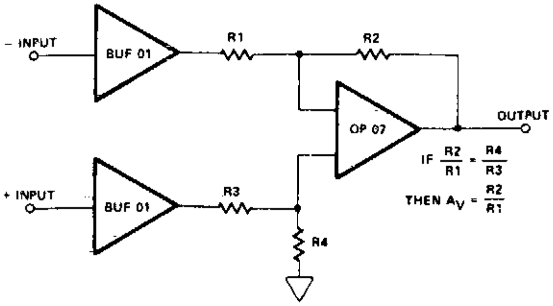

−
3
BUF-02
6
5
R2
500Ω
0.1%
R4
50kΩ
0.1%
1
V+
INPUTS
BALANCE
R1
100kΩ
2
OP-05
3
OUTPUT
5
1
+
3
BUF-02
6
R3
500Ω
0.1%
R5
50kΩ
0.1%

$\frac{R4}{R2} = \frac{R5}{R3}$

$A_V = \frac{R4}{R2}$

Fig. 43-18

HIGH IMPEDANCE DIFFERENTIAL AMPLIFIER

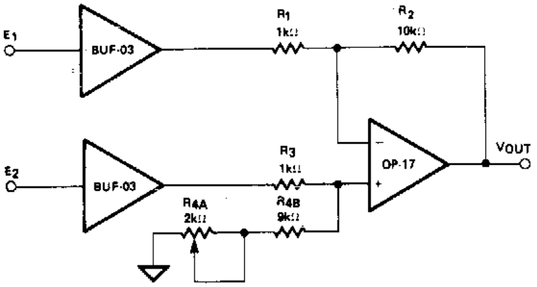

− INPUT
BUF 01
R1
R2
OP 07
OUTPUT
+ INPUT
BUF 01
R3
R4

IF $\frac{R2}{R1} = \frac{R4}{R3}$

THEN $A_V = \frac{R2}{R1}$

Fig. 43-19

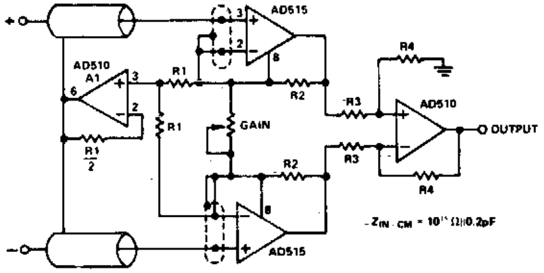
HIGH SPEED INSTRUMENTATION AMPLIFIER


E_1
BUF-03
R_1
1kΩ
R_2
10kΩ
−
OP-17
+
VOUT
E_2
BUF-03
R_3
1kΩ
R_{4A}
2kΩ
R_{4B}
9kΩ

NOTE: $\frac{R_2}{R_1} = \frac{R_4}{R_3}$ FOR GOOD COMMON MODE REJECTION.
R_{4A} IS ADJUSTED FOR BEST CMRR.

Fig. 43-20

VERY HIGH IMPEDANCE INSTRUMENTATION AMPLIFIER


+
3
AD515
2
8
R4
AD510
A1
6
3
2
R1
R2
R3
AD510
R1
GAIN
OUTPUT
$\frac{R1}{2}$
R2
R3
R4
8
−
AD515

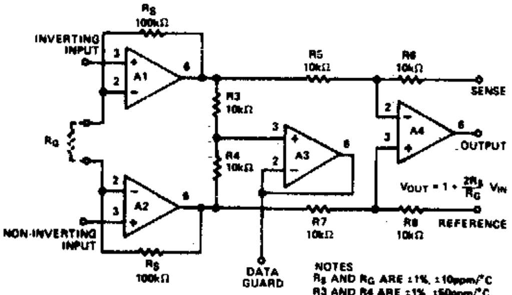
$-Z_{IN \cdot CM} = 10^{11}\ \Omega \| 0.2pF$

(ALL RESISTORS OF SAME NUMBER SHOULD BE MATCHED ±0.1%)

(BUFFER A1 BOOSTS COMMON MODE Z_{IN} BY DRIVING CABLE SHIELDS AT COMMON MODE VOLTAGE AND NEUTRALIZING CM CAPACITANCE)

Fig. 43-21

PRECISION FET INPUT INSTRUMENTATION AMPLIFIER


R_S
100kΩ
INVERTING INPUT
3
2
A1
6
R5
10kΩ
R6
10kΩ
SENSE
R3
10kΩ
R_G
3
A3
6
2
A4
6
3
OUTPUT
R4
10kΩ
2
2
A2
6
3
$V_{OUT} = 1 + \frac{2R_S}{R_G} V_{IN}$
R7
10kΩ
R8
10kΩ
REFERENCE
NON-INVERTING INPUT
R_S
100kΩ
DATA GUARD

NOTES
R_S AND R_G ARE ±1%, ±10ppm/°C
R3 AND R4 ARE ±1%, ±50ppm/°C.
R5, R6, R7, R8 ARE A MATCHED NETWORK, ±0.01%, ±2ppm/°C TRACKING TC.

ICs are AD547L.

Fig. 43-22

HIGH STABILITY THERMOCOUPLE AMPLIFIER

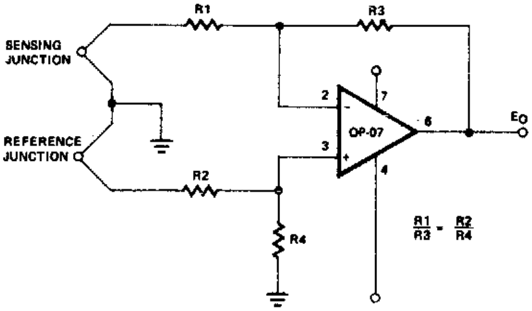

R1
R3
SENSING JUNCTION
2
7
6
E_O
OP-07
3
REFERENCE JUNCTION
R2
4
R4
$\frac{R1}{R3} = \frac{R2}{R4}$

Fig. 43-23

HIGH STABILITY THERMOCOUPLE AMPLIFIER

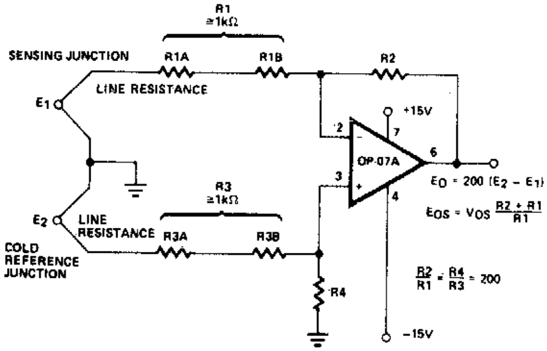
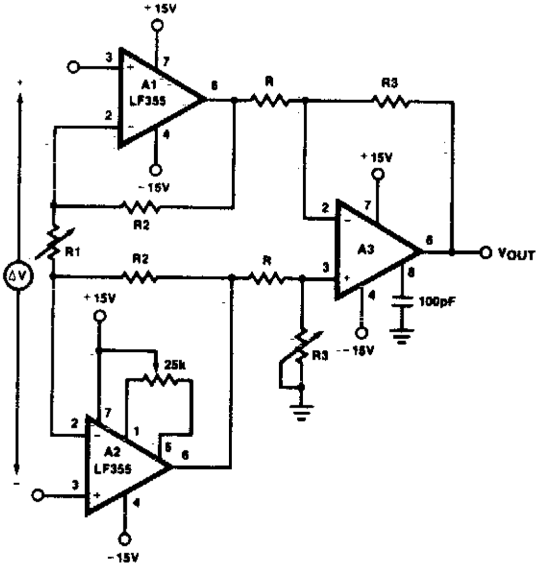

R1
≥1kΩ
SENSING JUNCTION
R1A
R1B
R2
LINE RESISTANCE
E_1
+15V
2
7
6
OP-07A
3
$E_O = 200\ (E_2 - E_1)$
4
R3
≥1kΩ
$E_{OS} = V_{OS} \frac{R2 + R1}{R1}$
E_2
LINE RESISTANCE
R3A
R3B
COLD REFERENCE JUNCTION
$\frac{R2}{R1} = \frac{R4}{R3} = 200$
R4
−15V

Fig. 43-24

HIGH IMPEDANCE LOW DRIFT INSTRUMENTATION AMPLIFIER


+15V
3
7
A1
LF355
6
R
R3
+
2
4
−15V
+15V
R2
2
7
R1
A3
6
VOUT
ΔV
R2
R
3
8
4
100pF
+15V
−15V
25k
R3
7
2
1
A2
LF355
5
6
−
3
4
−15V

- $V_{OUT} = \frac{R3}{R}\left[\frac{2R2}{R1} + 1\right]\Delta V$, V− +2V ≤ V_{IN} Common-Mode ≤ V+
- System V_{OS} adjusted via A2 V_{OS} adjust
- Trim R3 to boost up CMRR to 120dB.

Fig. 43-25

44

Light Activated Circuits

The sources of the following circuits are contained in the Sources section beginning on page 730. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

Pulse Generation by Interrupting a Light Beam
Optical Communication System
Four Quadrant Photo-Conductive Detector Amplifier
Precision Photodiode Comparator
Automatic Night Light
Receiver for 50 kHz FM Optional Transmitter
Photodiode Amplifier
Optical Schmitt Trigger
Adjustable Light Detection Switch
Photocell Memory Switch for AC Power Control
Optical Transmitter
Light Interruption Detector
Optical Receiver
Light Isolated Power Relay Circuit
Precision Photodiode Level Detector
Light Beam Operated On-Off Relay
Logarithmic Light Sensor
FM (PRM) Optical Transmitter
Light Level Sensor

PULSE GENERATION BY INTERRUPTING A LIGHT BEAM

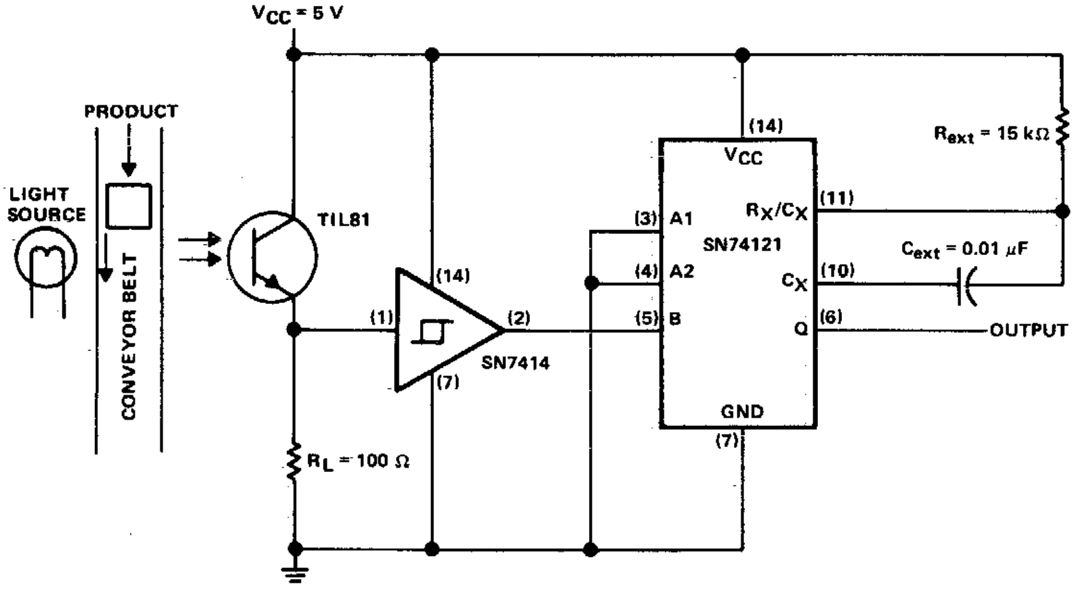
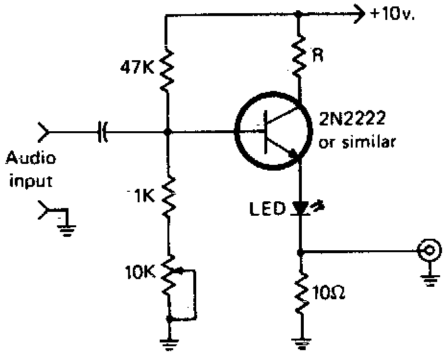

VCC = 5 V
PRODUCT
LIGHT SOURCE
CONVEYOR BELT
TIL81
(14)
(1)
(2)
SN7414
(7)
RL = 100 Ω
(14)
VCC
(3) A1
(4) A2
(5) B
SN74121
RX/CX (11)
CX (10)
Q (6)
OUTPUT
GND
(7)
Rext = 15 kΩ
Cext = 0.01 µF

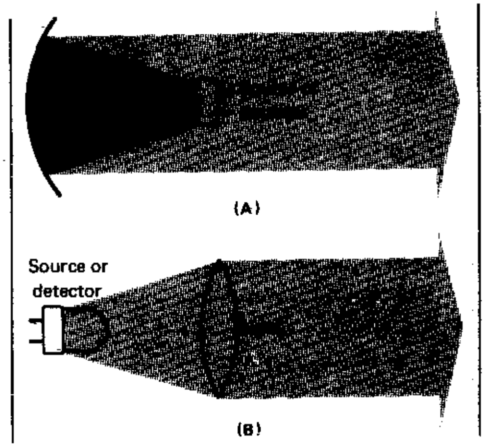
Fig. 44-1

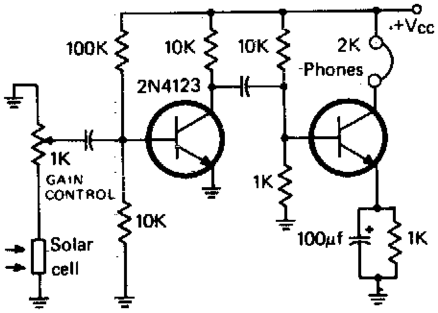
Circuit Notes

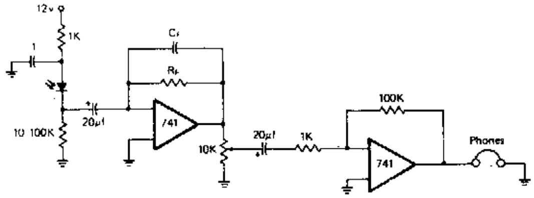
This circuit puts out a pulse when an object on the conveyor belt blocks the light source. The light source keeps the phototransistor turned on. This produces a high-logic-level voltage at the Schmitt-trigger inverter and a TTL-compatible low logic level at pin 5 of the monostable. When an object blocks the light, TIL81 turns off the Schmitt-trigger inverter to triggers the one shot.

OPTICAL COMMUNICATION SYSTEM


+10v.
47K
R
2N2222
or similar
Audio
input
1K
LED
10K
10Ω


(A)
Source or
detector
(B)


100K
10K
10K
2K
+Vcc
Phones
2N4123
1K
GAIN
CONTROL
10K
Solar
cell
1K
100µf
1K

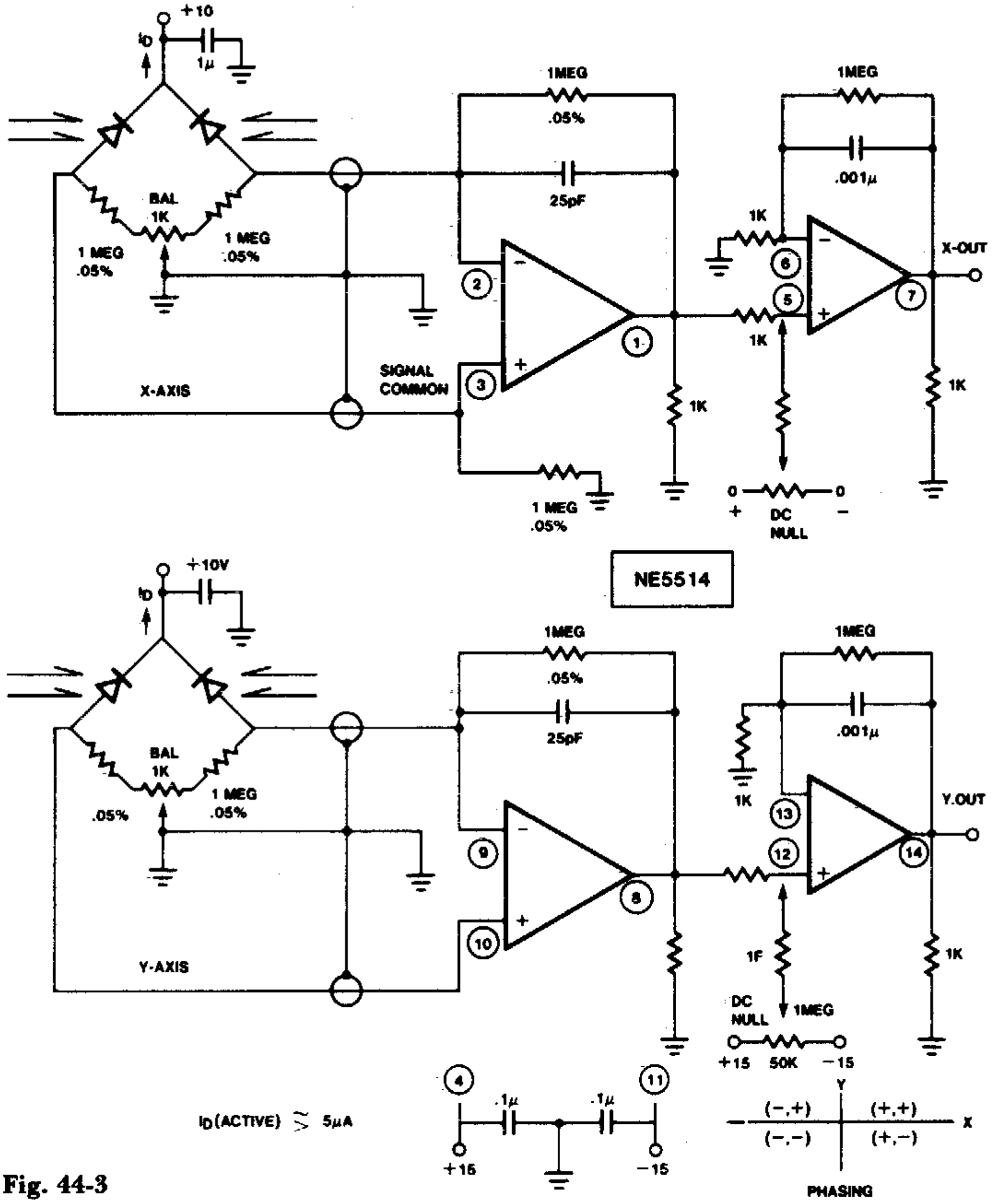

12v
1K
1
Cf
Rf
20µf
741
10 100K
10K
20µf
1K
100K
741
Phones

Circuit Notes

The simple modulator stage will accommodate most common LEDs. By adjusting the potentiometer, the bias of the transistor is varied until the LED is at its half output point. Then, audio will cause it to vary above and below this point. The purpose of R1 is to limit the current through the LED to a safe level and the purpose of the 10 ohm resistor is to allow a portion of the modulating signal to be observed on a scope.

Fig. 44-2

FOUR QUADRANT PHOTO-CONDUCTIVE DETECTOR AMPLIFIER



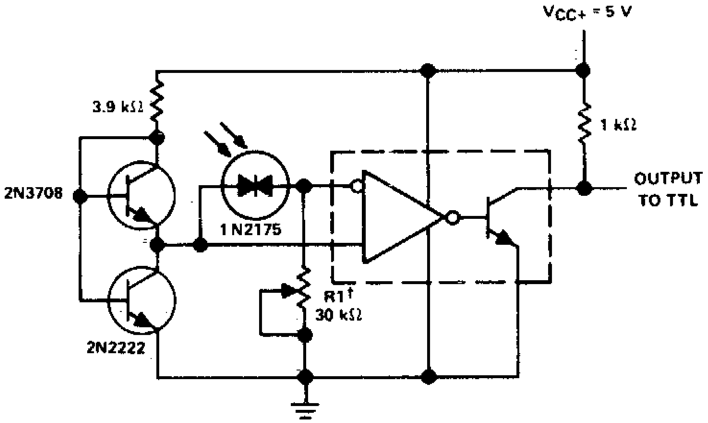
+10
I_D
1μ
1MEG
.05%
25pF
1MEG
.001μ
1K
X-OUT
BAL
1K
1 MEG
.05%
1 MEG
.05%
2
3
1
6
5
7
SIGNAL
COMMON
X-AXIS
1K
1K
1K
1 MEG
.05%
+
DC
NULL
−
NE5514
+10V
I_D
1MEG
.05%
25pF
1MEG
.001μ
BAL
1K
1 MEG
.05%
.05%
1K
Y.OUT
9
10
8
13
12
14
Y-AXIS
1F
1K
DC
NULL
1MEG
+15
50K
−15
4
11
.1μ
.1μ
+15
−15
I_D(ACTIVE) ≳ 5μA
Y
(−,+)
(+,+)
(−,−)
(+,−)
X
PHASING

Fig. 44-3

Circuit Notes

Use this circuit to sense four quadrant motion of a light source. By proper summing of the signals from the X and Y axes, four quadrant output may be fed to an X-Y plotter, oscilloscope, or computer for simulation. IC = NE/SE5514

PRECISION PHOTODIODE COMPARATOR



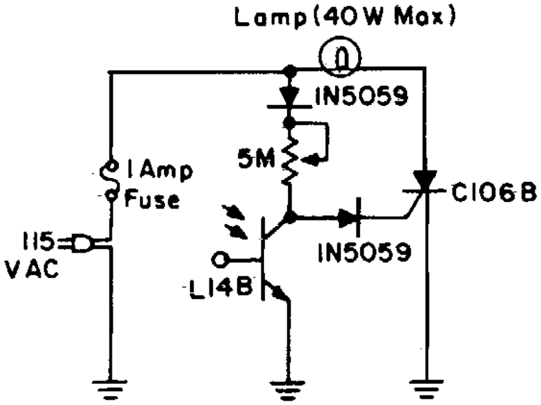
Vcc+ = 5 V
3.9 kΩ
1 kΩ
2N3708
1N2175
OUTPUT TO TTL
R1†
30 kΩ
2N2222

Fig. 44-4

Circuit Notes

R1 sets the comparison level. At comparison, the photodiode has less than 5 mV across it, decreasing dark current by an order of magnitude. IC = LM 111/211/311.

AUTOMATIC NIGHT LIGHT



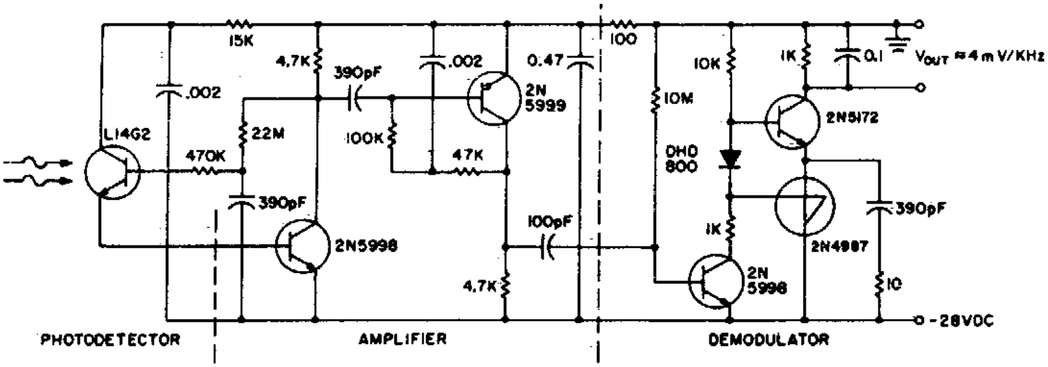
Lamp (40 W Max)
1N5059
1 Amp Fuse
5M
C106B
115 VAC
1N5059
L14B

Fig. 44-5

Circuit Notes

During daylight hours, the L14B photo-Darlington (JEDEC registered as 2N5777 through 2N5780) shunts all gate current to ground. At night, the L14B effectively provides a high resistance, diverting the current into the gate of the C106B and turning on the lamp.

RECEIVER FOR 50 kHz FM OPTICAL TRANSMITTER



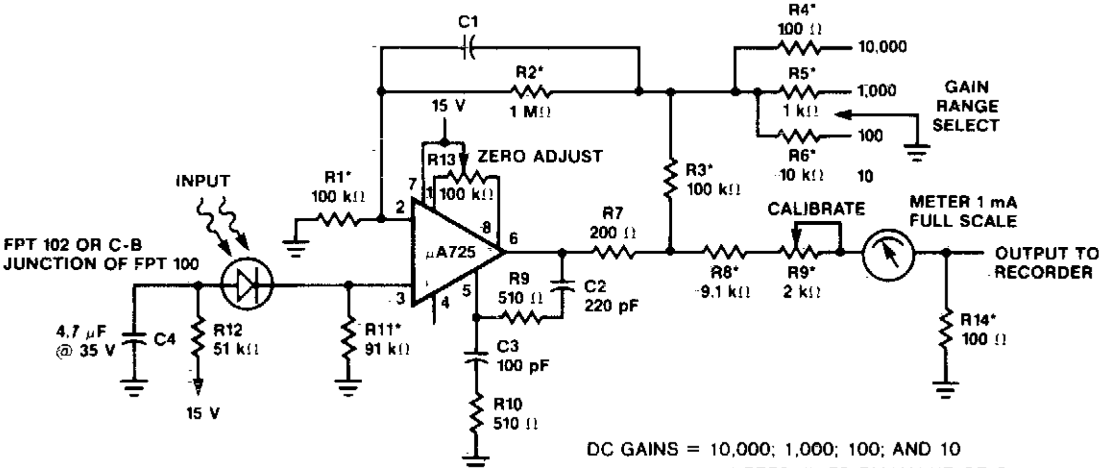
15K
4.7K
390pF
.002
0.47
100
10K
1K
0.1
V_OUT ≈4 mV/KHz
.002
2N 5999
10M
2N5172
L14G2
22M
100K
470K
47K
DHD 800
390pF
390pF
100pF
1K
2N5998
2N4987
4.7K
2N 5998
10
-28VDC
PHOTODETECTOR
AMPLIFIER
DEMODULATOR

Fig. 44-6

Circuit Notes

This circuit consists of a L14G2 detector, two stages of gain, and a FM demodulator. Better sensitivity can be obtained using more stages of stabilized gain with AGC.

PHOTODIODE AMPLIFIER



C1
R4*
100 Ω
10,000
R2*
R5*
1,000
GAIN
RANGE
SELECT
15 V
1 MΩ
1 kΩ
100
R6*
R13
ZERO ADJUST
10 kΩ
10
R3*
100 kΩ
INPUT
R1*
100 kΩ
7
1
100 kΩ
2
8
R7
200 Ω
CALIBRATE
METER 1 mA
FULL SCALE
FPT 102 OR C-B
JUNCTION OF FPT 100
μA725
6
OUTPUT TO
RECORDER
R8*
9.1 kΩ
R9*
2 kΩ
R9
510 Ω
C2
220 pF
3
4
5
4.7 μF
@ 35 V
C4
R12
51 kΩ
R11*
91 kΩ
R14*
100 Ω
C3
100 pF
15 V
R10
510 Ω

DC GAINS = 10,000; 1,000; 100; AND 10
BANDWIDTH = DETERMINED BY VALUE OF C_1

Fig. 44-7

OPTICAL SCHMITT TRIGGER

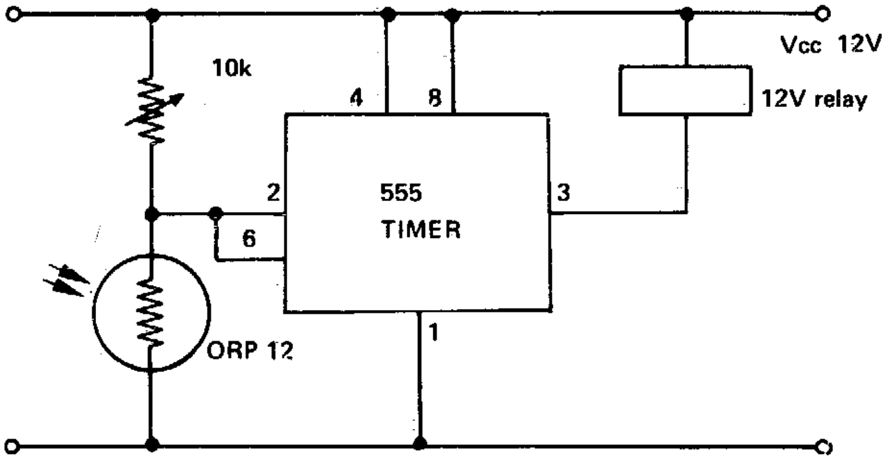

Vcc 12V
10k
4
8
12V relay
2
555
TIMER
3
6
1
ORP 12

Fig. 44-8

Circuit Notes

This circuit shows a 555 with its trigger and threshold inputs connected together used to energize a relay when the light level on a photoconductive cell falls below a preset value. Circuit can be used in other applications where a high input impedance and low output impedance are required with the minimum component count.

ADJUSTABLE LIGHT DETECTION SWITCH

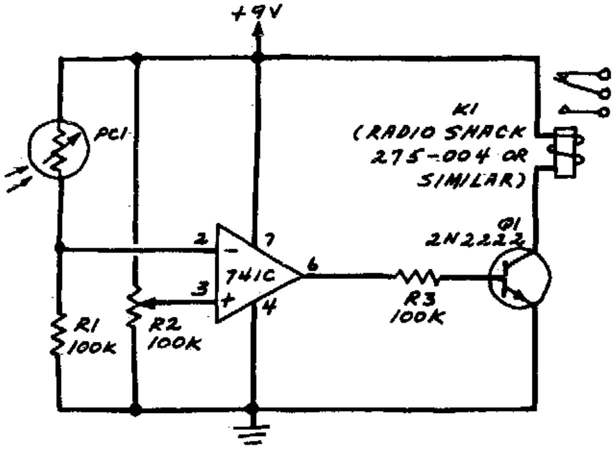
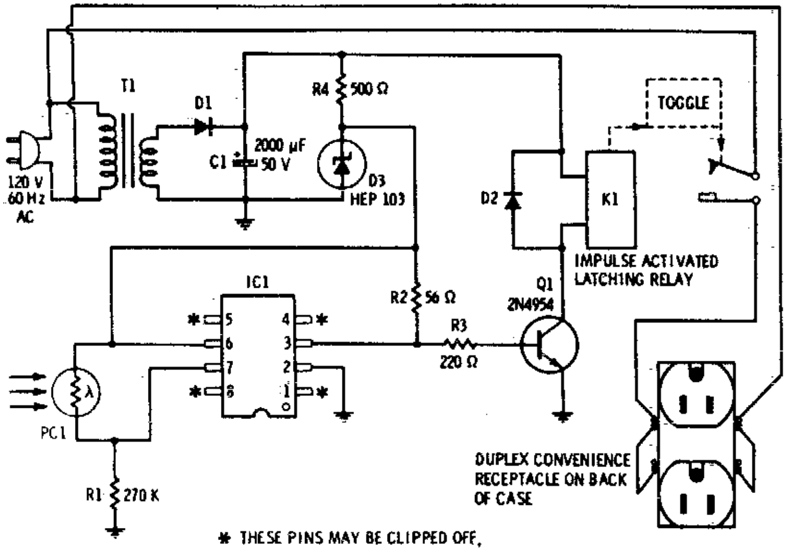

+9V
PC1
K1
(RADIO SHACK
275-004 OR
SIMILAR)
Q1
2N2222
2
7
741C
6
3
4
R3
100K
R1
100K
R2
100K

Fig. 44-9

Circuit Notes

R2 sets the circuit's threshold. When the light intensity at PCI's surface is decreased, the resistance of PC1 a cadmium-sulfide photoresistor is increased. This decreases the voltage at the inverting input of the 741. When the reference voltage at the 741's noninverting input is properly adjusted via R2, the comparator will switch from low to high when PC1 is darkened. This turns on Q1 which, in turn, pulls in relay K1.

PHOTOCELL MEMORY SWITCH FOR AC POWER CONTROL


T1
D1
R4
500 Ω
2000 µF
50 V
C1
D3
HEP 103
120 V
60 Hz
AC
TOGGLE
D2
K1
IMPULSE ACTIVATED
LATCHING RELAY
IC1
R2
56 Ω
Q1
2N4954
R3
220 Ω
5
4
6
3
7
2
8
1
PC1
R1
270 K
DUPLEX CONVENIENCE
RECEPTACLE ON BACK
OF CASE

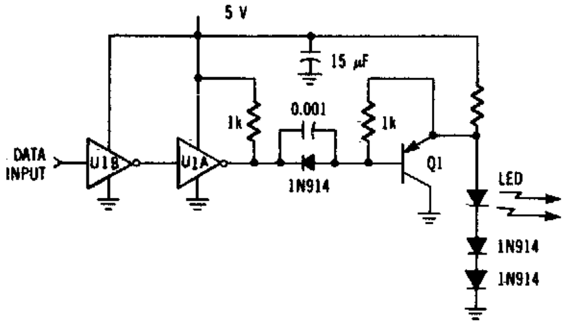
* THESE PINS MAY BE CLIPPED OFF, IF DESIRED.

Fig. 44-10

Circuit Notes

Provides remote control for ac-powered devices by using the beam of a flashlight as a magic wand. The important aspect of this gadget is that it remembers. Activate it once to apply power to a device and it stays on. Activate it a second time and power goes off and stays off. It consists of a combination of a high-sensitivity photocell, a high-gain IC Schmitt trigger, and an impulse-actuated latching relay.

OPTICAL TRANSMITTER


5 V
15 µF
1k
0.001
1k
DATA
INPUT
U1B
U1A
1N914
Q1
LED
1N914
1N914

Circuit Notes

Driver circuit uses an MC74LS04 and one discrete transistor. The circuit can drive the LED (MFOE1200) at up to 1 Mbps data rate.

Fig. 44-11

LIGHT INTERRUPTION DETECTOR

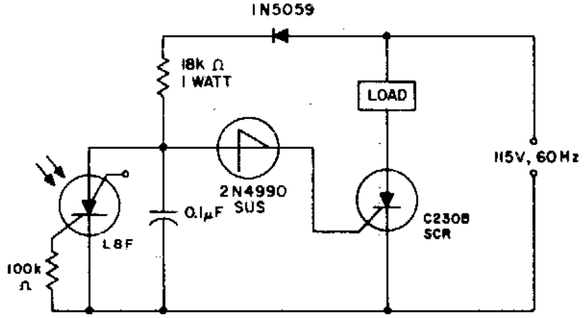
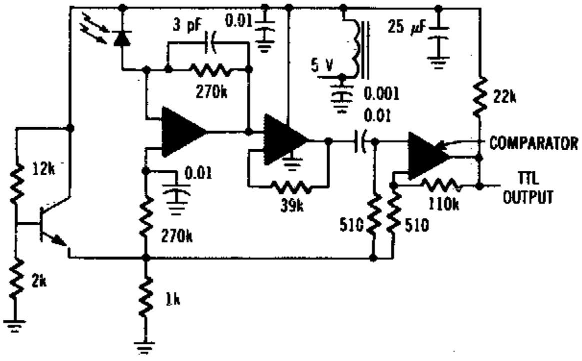

1N5059
18k Ω
1 WATT
LOAD
115V, 60Hz
2N4990
SUS
0.1μF
L8F
100k
Ω
C230B
SCR

Fig. 44-12

Circuit Notes

When the light incident on the LASCR is interrupted, the voltage at the anode to the 2N4990 unilateral switch goes positive on the next positive cycle of the power which in turn triggers the switch and the C230 SCR when the switching voltage of the unilateral switch is reached. This will cause the load to be energized for as long as light is not incident on the LASCR.

OPTICAL RECEIVER


3 pF
0.01
25 μF
5 V
270k
0.001
0.01
22k
COMPARATOR
12k
0.01
39k
110k
TTL
OUTPUT
510
510
270k
2k
1k

Circuit Notes

The MFOD1100 PIN diode requires shielding from emi.

Fig. 44-13

LIGHT ISOLATED SOLID STATE POWER RELAY CIRCUITS

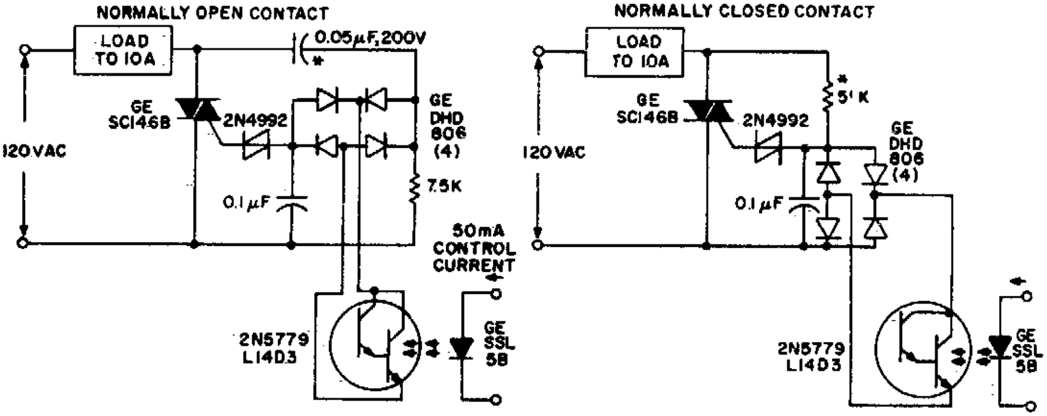
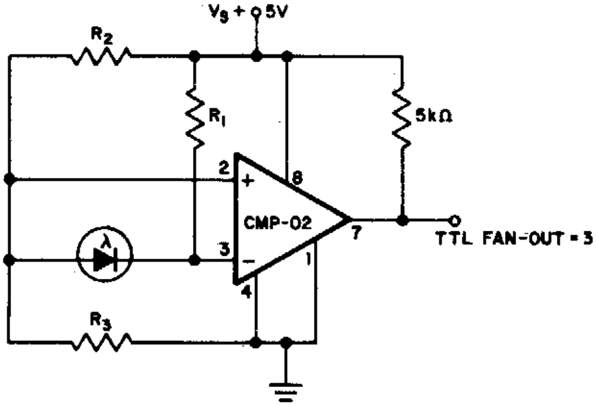

NORMALLY OPEN CONTACT
LOAD TO 10A
0.05μF,200V
*
GE SC146B
2N4992
GE DHD 806 (4)
120VAC
7.5K
0.1μF
50mA CONTROL CURRENT
2N5779 L14D3
GE SSL 5B
NORMALLY CLOSED CONTACT
LOAD TO 10A
*
5.1K
GE SC146B
2N4992
GE DHD 806 (4)
120 VAC
0.1μF
2N5779 L14D3
GE SSL 5B

Fig. 44-14

Circuit Notes

Both circuits use the G.E. SC146B, 200 V, 10 A Triac as load current contacts. These triacs are triggered by normal SBS (2N4992) trigger circuits, which are controlled by the photo-Darlington, acting through the DA806 bridge as an ac photo switch. To operate the relays at other line voltages the asterisked (*) components are scaled to supply identical current. Ratings must be changed as required. Incandescent lamps may be used in place of the light emitting diodes, if desired.

PRECISION PHOTODIODE LEVEL DETECTOR

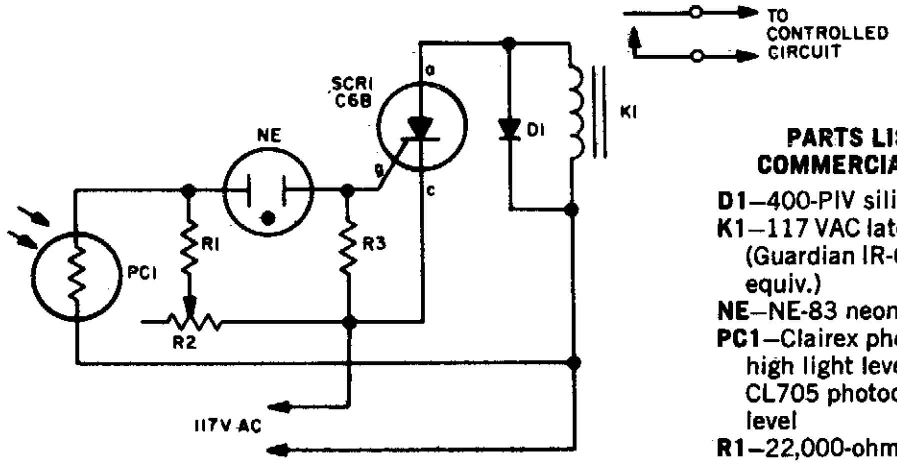

V_S + 5V
R_2
R_1
5kΩ
2
+
8
CMP-02
7
TTL FAN-OUT = 3
λ
3
−
1
4
R_3

Circuit Notes

For R1 = 2.5 M, R2 = R3 = 5 M. The output state changes at a photo diode current of 0.5 μA.

Fig. 44-15

LIGHT BEAM OPERATED ON-OFF RELAY


TO
CONTROLLED
CIRCUIT
SCR1
C6B
a
g
c
NE
D1
K1
PC1
R1
R3
R2
117V AC

PARTS LIST FOR COMMERCIAL KILLER

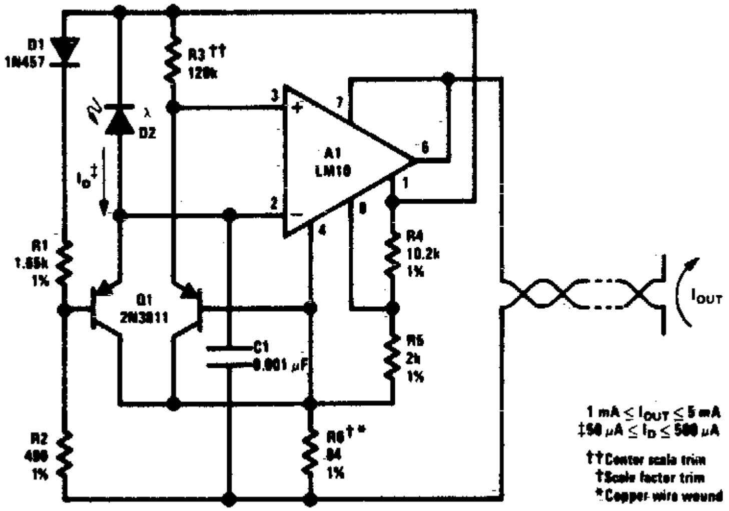
D1—400-PIV silicon rectifier
K1—117 VAC latching relay (Guardian IR-610L-A115 or equiv.)
NE—NE-83 neon lamp
PC1—Clairex photo cell CL505 for high light level; CL704 or CL705 photocell for low light level
R1—22,000-ohm, ½-watt resistor
R2—1-megohm potentiometer
R3—100-ohm, ½-watt resistor
SCR1—HEP R1218, 200V, 4A, silicon-controlled rectifier

Fig. 44-16

Circuit Notes

When a beam of light strikes the photocell, the voltage across neon lamp NE-1 rises sharply. NE-1 turns on and fires the SCR. K1 is an impulse relay whose contacts stay in position even after coil current is removed. The first impulse opens K1's contacts, the second impulse closes them, etc.

LOGARITHMIC LIGHT SENSOR


D1
1N457
R3 ††
120k
λ
D2
I_D ‡
3
+
7
A1
LM10
6
1
2
−
8
4
R1
1.65k
1%
Q1
2N3811
R4
10.2k
1%
C1
0.001 μF
R5
2k
1%
R2
499
1%
R6 †*
84
1%
I_{OUT}

$1\ mA \le I_{OUT} \le 5\ mA$
‡$50\ \mu A \le I_D \le 500\ \mu A$

††Center scale trim
†Scale factor trim
*Copper wire wound

Fig. 44-17

FM (PRM) OPTICAL TRANSMITTER

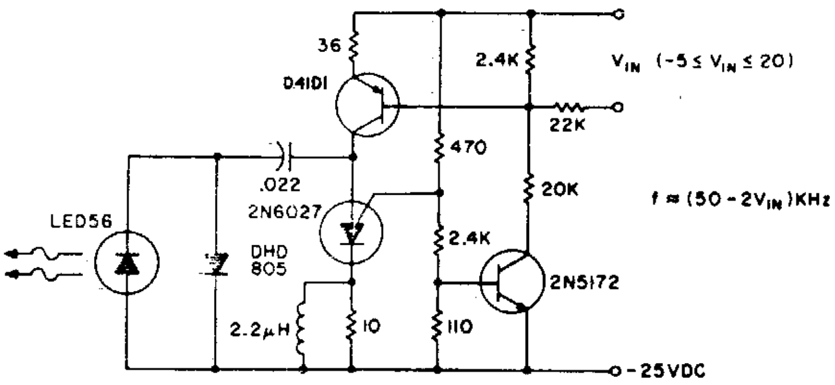
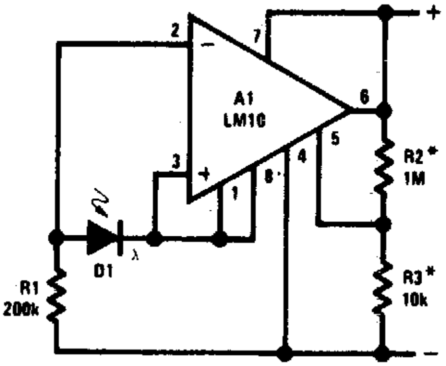

36
D41D1
2.4K
V_{IN} $(-5 \le V_{IN} \le 20)$
22K
470
.022
2N6027
LED56
DHD
805
20K
$f \approx (50 - 2V_{IN})$KHz
2.4K
2N5172
2.2µH
10
110
-25VDC

Fig. 44-18

Circuit Notes

The basic circuit can be operated at 80 kHz and is limited by the PUT capacitor combination. 60 kHz is the maximum modulation frequency. The pulse repetition rate is a linear function of V_{IN}, the modulating voltage. Lenses or reflectors minimizes stray light noise effects. Greater output can be obtained by using a larger capacitor, which also gives a lower operating frequency, or using a higher power output IRED such as the F5D1. Average power consumption of the transmitter circuit is less than 3 watts.

LIGHT LEVEL SENSOR


+
2
7
A1
LM10
6
5
3
4
8
1
R2*
1M
D1
λ
R1
200k
R3*
10k
-

* provides hysteresis

Fig. 44-19

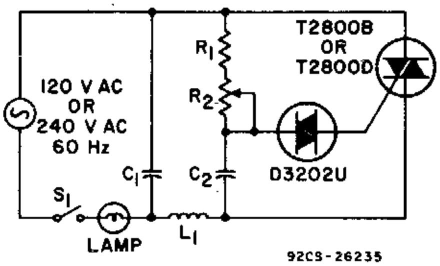
45

Light Controls

The sources of the following circuits are contained in the Sources section beginning on page 730. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

Light Dimmers
Remote Control for Lamp or Appliance
High Power Control for Sensitive Contacts
Complementary Lighting Control
Floodlamp Power Control
Hysteresis-Free Phase Control Circuit
Low Cost Lamp Dimmer
Zero Point Switch
800 W Triac Light Dimmer
Full-Wave SCR Control
860 W Limited Range Low Cost Precision Light Control
800 W Soft-Start Light Dimmer
Low Loss Brightness Control
Half-Wave Ac Phase-Controlled Circuit
Emergency Light
Neon Lamp Driver
Complementary Ac Power Switching
Battery Lantern Circuit
Shift Register
Light Level Controller
2.2 W Incandescent Lamp Driver

LIGHT DIMMERS



(a) Single-time-constant light-dimmer circuit.

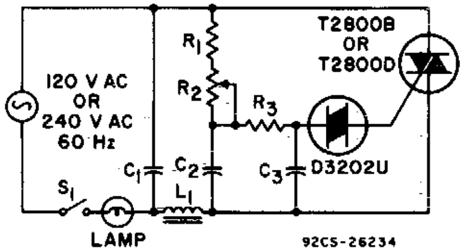
Parts List

120-Volt, 60-Hz Operation

C_1, C_2 = 0.1 μF, 200 V
L_1 = 100 μH
R_1 = 3300 ohms, 0.5 watt
R_2 = light control, potentiometer, 0.25 megohm, 0.5 watt

240-Volt, 50/60 Hz Operation

C_1 = 0.1 μF, 400 V
C_2 = 0.05 μF, 400 V
L_1 = 200 μH
R_1 = 4700 ohms, 0.5 watt
R_2 = light control, potentiometer, 0.25 megohm, 1 watt



(b) Double-time-constant light-dimmer circuit.

Parts List

120-Volt, 60-Hz Operation

C_1, C_2 = 0.1 μF, 200 V
C_3 = 0.1 μF, 100 V
L_1 = 100 μH
R_1 = 1000 ohms, 0.5 watt
R_2 = light control, potentiometer, 0.1 megohm, 0.5 watt

240-Volt, 60-Hz Operation

C_1 = 0.1 μF, 400 V
C_2 = 0.05 μF, 400 V
C_3 = 0.1 μF, 100 V
L_1 = 100 μH
R_1 = 7500 ohms, 2 watts
R_2 = light control, potentiometer, 0.2 megohm, 1 watt
R_3 = 7500 ohms, 2 watts

Fig. 45-1

Circuit Notes

The two lamp-dimmer circuits differ in that (a) employs a single-time-constant trigger network and (b) uses a double-time-constant trigger circuit that reduces hysteresis effects and thereby extends the effective range of the light-control potentiometer. (Hysteresis refers to a difference in the control potentiometer setting at which the lamp turns on and the setting at which the light is extinguished.) The additional capacitor C2 in (b) reduces hysteresis by charging to a higher voltage than capacitor C3. During gate triggering, C3 discharges to form the gate current pulse. Capacitor C2, however, has a longer discharge time constant and this capacitor restores some of the charge removed from C3 by the gate current pulse.

REMOTE CONTROL FOR LAMP OR APPLIANCE

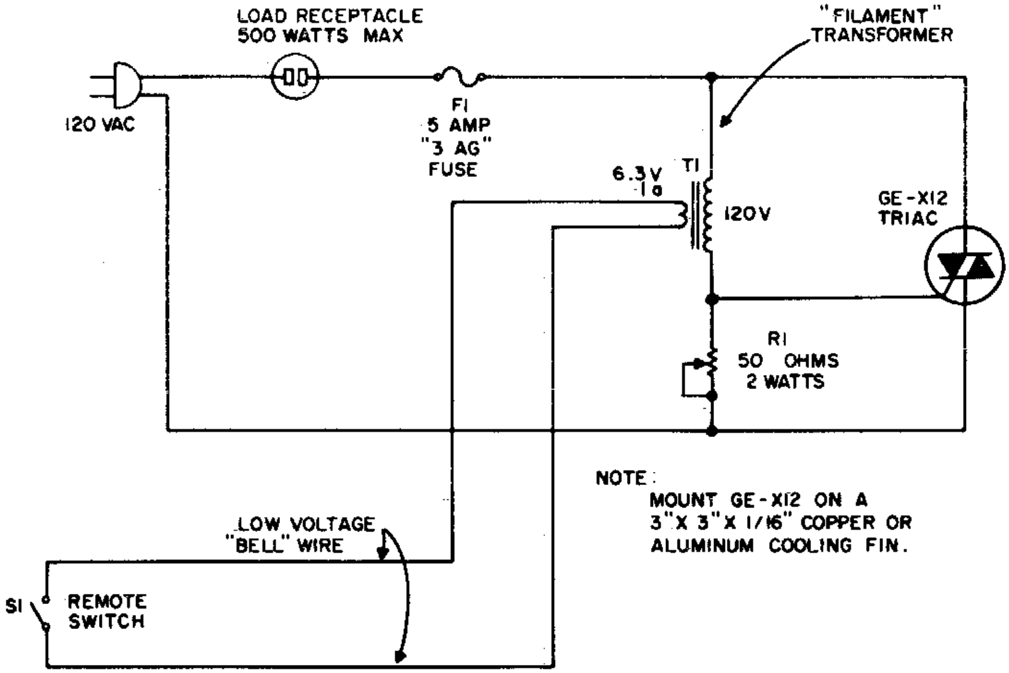

LOAD RECEPTACLE
500 WATTS MAX
"FILAMENT"
TRANSFORMER
120 VAC
F1
5 AMP
"3 AG"
FUSE
6.3V
1 a
T1
120V
GE-X12
TRIAC
R1
50 OHMS
2 WATTS
NOTE:
MOUNT GE-X12 ON A
3" X 3" X 1/16" COPPER OR
ALUMINUM COOLING FIN.
LOW VOLTAGE
"BELL" WIRE
S1
REMOTE
SWITCH

Fig. 45-2

Circuit Notes

The circuit uses the primary current of a small 6.3 volt filament transformer to actuate a triac and energize the load. When switch S1, in the six-volt secondary, of the transformer is open, a small "magnetizing" current flows through the primary winding. This magnetizing current may be large enough to trigger the triac. Therefore, a shunting resistor, R1, is required to prevent such triggering. R1, is adjusted for the highest resistance that will not cause the triac to trigger with S1 open. When single-pole remote switch, S1, closes, the secondary of the transformer is shorted and a high current flows through the 120-volt primary. This triggers the triac and energizes the load. When the triac conducts, current through the primary stops and thus prevents burning out the transformer.

HIGH POWER CONTROL FOR SENSITIVE CONTACTS

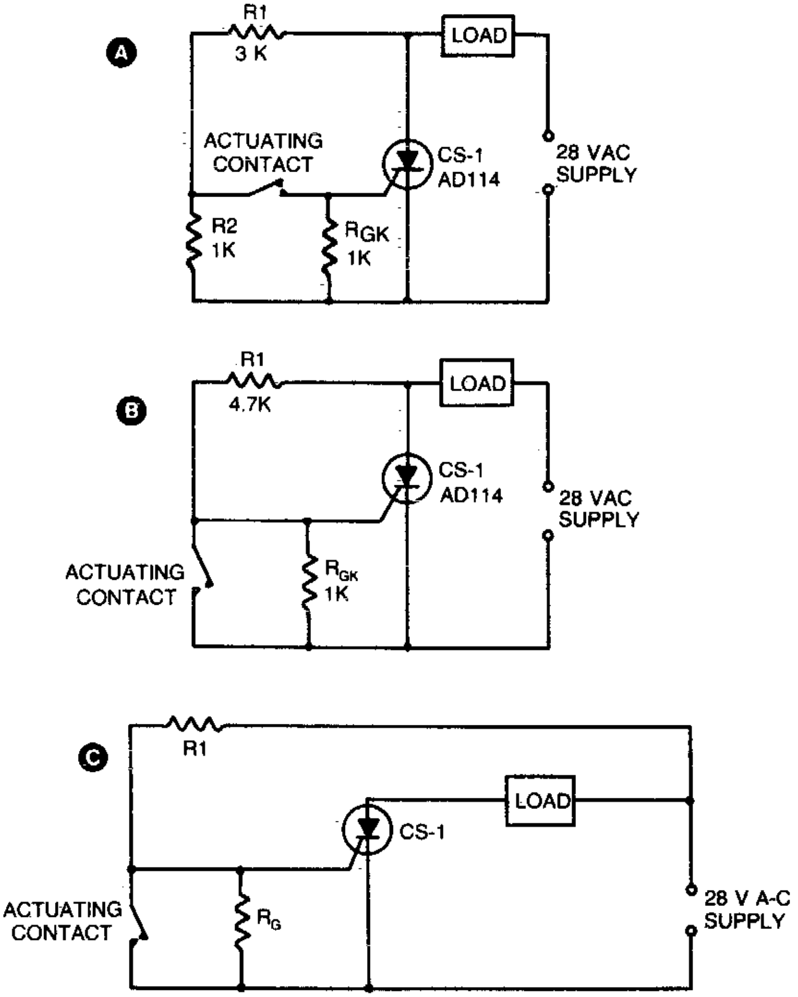

A
R1
3 K
LOAD
ACTUATING CONTACT
CS-1
AD114
28 VAC SUPPLY
R2
1K
RGK
1K
B
R1
4.7K
LOAD
CS-1
AD114
28 VAC SUPPLY
ACTUATING CONTACT
RGK
1K
C
R1
LOAD
CS-1
ACTUATING CONTACT
RG
28 V A-C SUPPLY

Fig. 45-3

Circuit Notes

Two simple arrangements for resistive loads are shown in A & B. The circuit in A will provide load power when the actuating contact is closed, and no power when the contact is open. B provides the reverse of this action—power being supplied to the load when the contact is open with no load power when the contact is closed. If desired, both circuits can be made to latch by operating with dc instead of the indicated ac supply. In both of these circuits, voltage across the sensitive contacts is under 5 volts, and contact current is below 5 mA. For inductive loads, R1 would normally be returned to the opposite side of the load as shown in C.

COMPLEMENTARY LIGHTING CONTROL

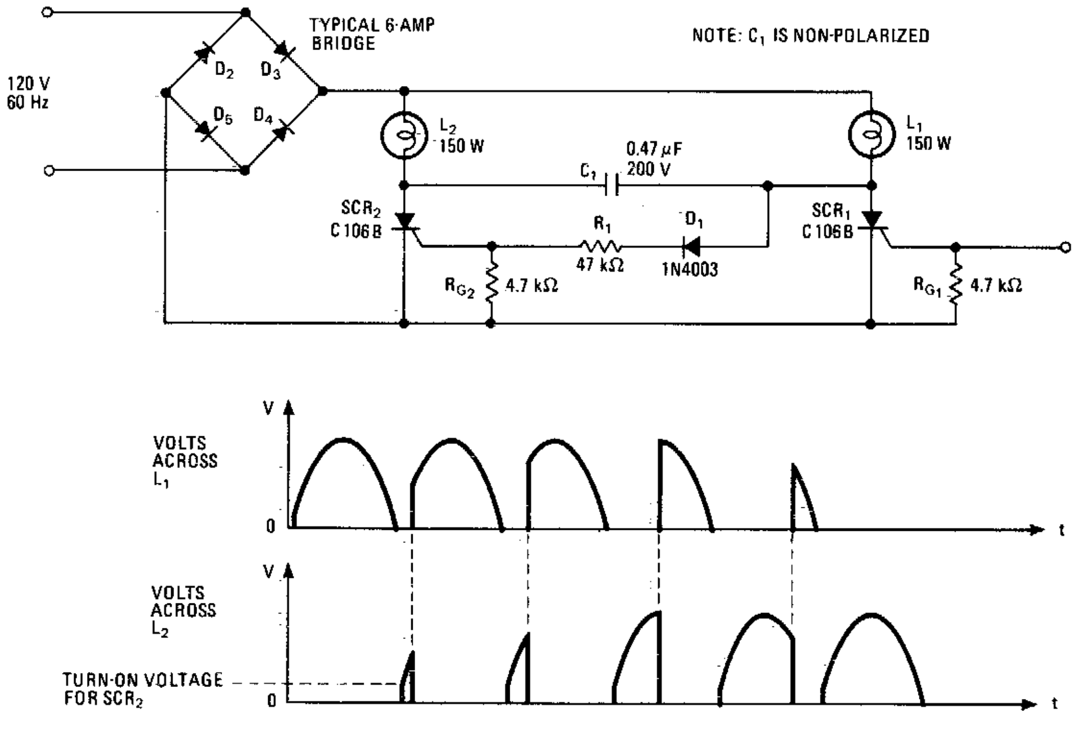

TYPICAL 6-AMP BRIDGE
NOTE: C_1 IS NON-POLARIZED
120 V
60 Hz
D_2
D_3
D_5
D_4
L_2
150 W
L_1
150 W
0.47 µF
C_1
200 V
SCR_2
C106B
SCR_1
C106B
R_1
47 kΩ
D_1
1N4003
R_{G2}
4.7 kΩ
R_{G1}
4.7 kΩ
V
VOLTS ACROSS L_1
0
t
V
VOLTS ACROSS L_2
TURN-ON VOLTAGE FOR SCR_2
0
t

Fig. 45-4

Circuit Notes

This lighting-control unit will fade out one lamp while simultaneously increasing the light output of another. The two loads track each other accurately without adjustments. The gate of SCR1, a silicon-controlled rectifier, is driven from a standard phase-control circuit, based, for example, on a unijunction transistor or a diac. It controls the brightness of lamp L1 directly. Whenever SCR1 is not on, a small current flows through L1, D1, and R1, permitting SCR2 to fire. When SCR1 turns on, current flow ceases through D1 and R1; the energy stored in C1 produces a negative spike that turns SCR2 off.

FLOODLAMP POWER CONTROL

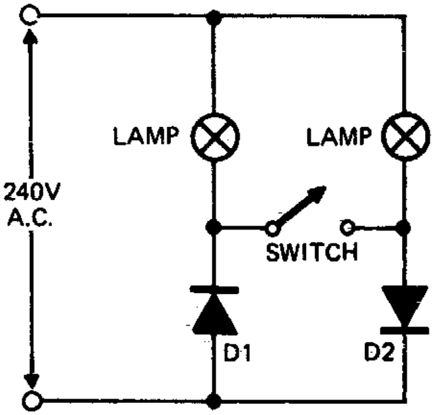

LAMP
LAMP
240V
A.C.
SWITCH
D1
D2

Fig. 45-5

Circuit Notes

When setting up photographic floodlamps, it is sometimes desirable to operate the lamps at lower power levels until actually ready to take the photograph. The circuit allows the lamps to operate on half cycle power when the switch is open, and full power, when the switch is closed. The diodes D1 and D2 should have a 400 volt PIV rating at 5 amps.

HYSTERESIS-FREE PHASE CONTROL CIRCUIT

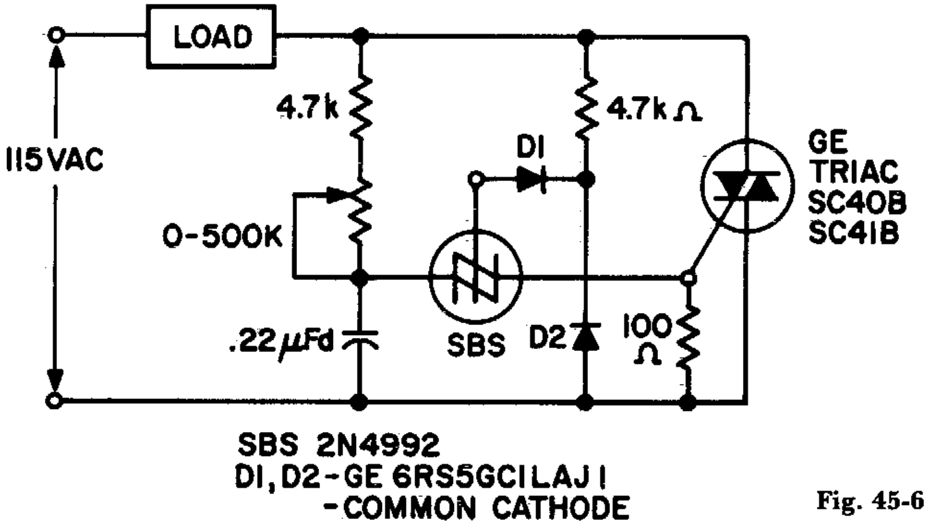

LOAD
115VAC
4.7k
4.7kΩ
0-500K
.22μFd
D1
SBS
D2
100Ω
GE
TRIAC
SC40B
SC41B
SBS 2N4992
D1, D2-GE 6RS5GC1LAJ1
-COMMON CATHODE

Fig. 45-6

Circuit Notes

This circuit is intended for lamp dimming and similar applications. It requires only one RC phase lag network. To avoid the hysteresis (or "snap-on") effect, the capacitor is reset to approximately 0 volts at the end of every positive half cycle using the gate lead.

LOW COST LAMP DIMMER

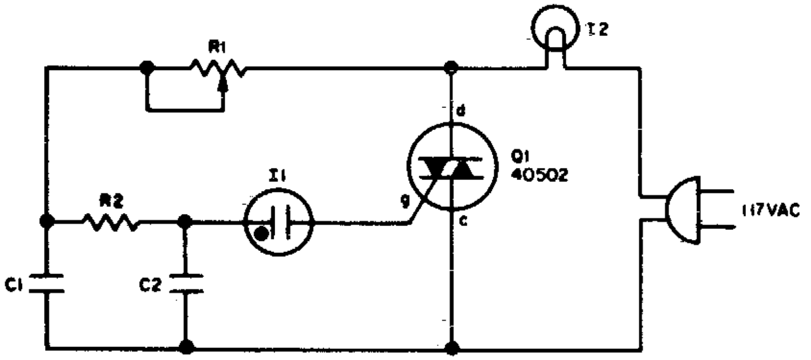

R1
I2
d
Q1
40502
I1
R2
g
c
117VAC
C1
C2

Fig. 45-7

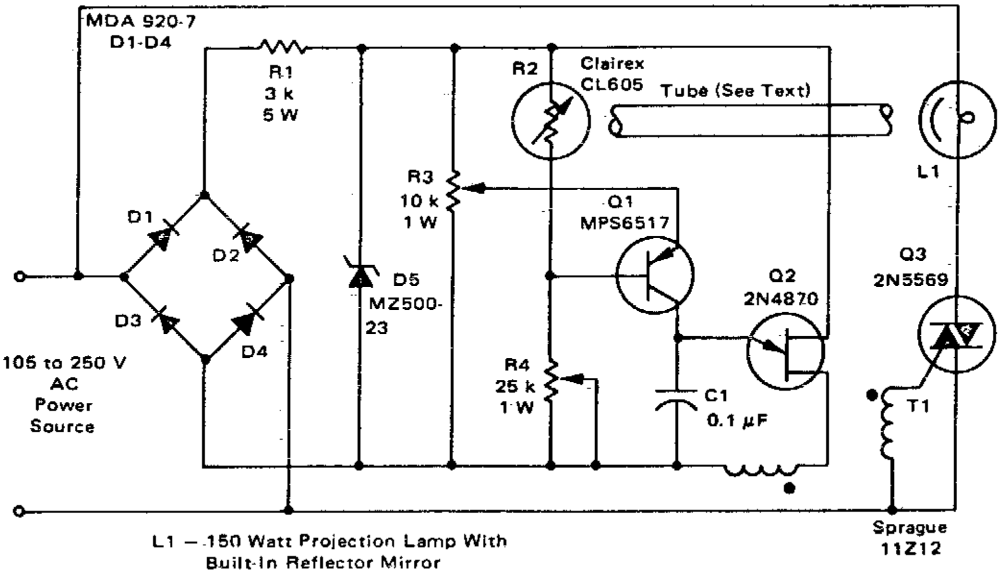
PARTS LIST FOR LO-COST LAMP DIMMER

C1, C2—0.068-uF, 200-VDC capacitor
I1—NE-2 neon lamp
I2—External lamp not to exceed 400 watts
Q1—RCA 40502 Triac
R1—50,000-ohm, pot.
R2—15,000-ohm, ½-watt resistor

Circuit Notes

Without a heatsink, Triac Q1 handles up to a 400-watt lamp. The neon lamp does not trip the gate until it conducts so the lamp turns on a medium brilliance. The lamp can then be backed off to a soft glow.

ZERO-POINT SWITCH


MDA 920-7
D1-D4
R1
3 k
5 W
R2
Clairex
CL605
Tube (See Text)
L1
R3
10 k
1 W
Q1
MPS6517
D1
D2
D5
MZ500-
23
Q2
2N4870
Q3
2N5569
D3
D4
105 to 250 V
AC
Power
Source
R4
25 k
1 W
C1
0.1 µF
T1
Sprague
11Z12

L1 — 150 Watt Projection Lamp With Built-In Reflector Mirror

Fig. 45-8

800 W TRIAC LIGHT DIMMER

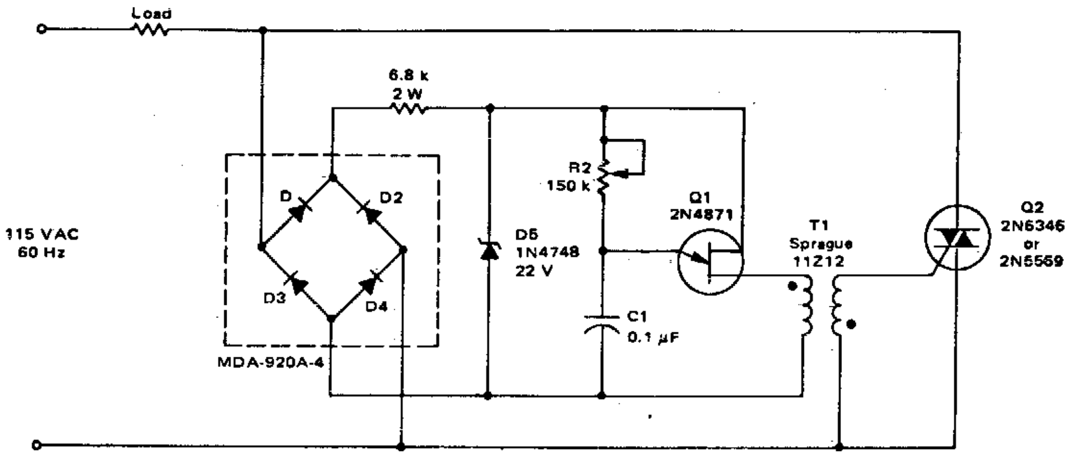

Load
6.8 k
2 W
R2
150 k
Q1
2N4871
D
D2
D3
D4
MDA-920A-4
D5
1N4748
22 V
C1
0.1 μF
115 VAC
60 Hz
T1
Sprague
11Z12
Q2
2N6346
or
2N5569

Fig. 45-9

FULL-WAVE SCR CONTROL

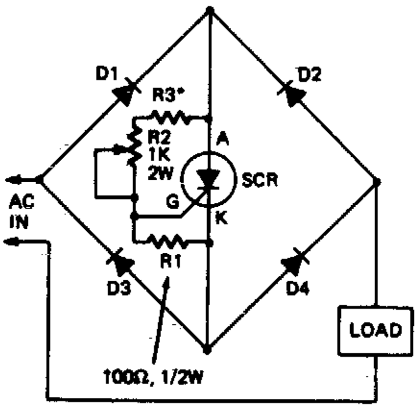
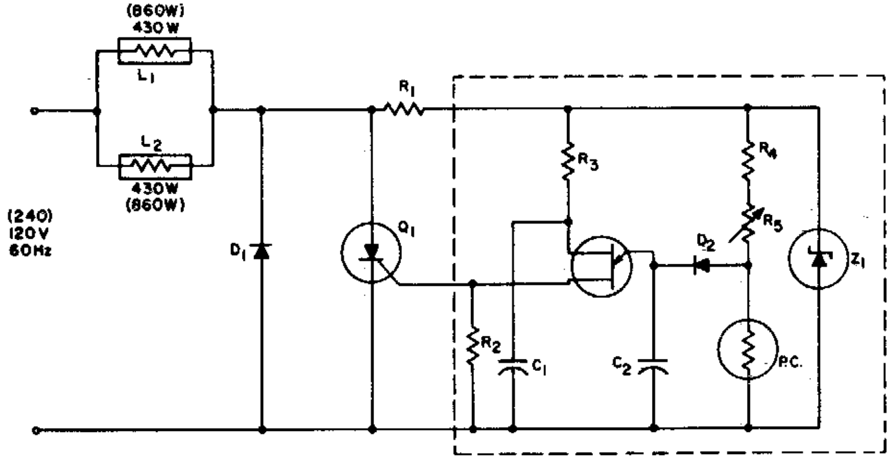

D1
R3*
D2
R2
1K
2W
A
SCR
G
K
AC
IN
R1
D3
D4
LOAD
100Ω, 1/2W

Fig. 45-10

Circuit Notes

This circuit enables a single SCR to provide fullwave control of resistive loads. Resistor R3 should be chosen so that when potentiometer R2 is at its minimum setting, the current in the load is at the required minimum level. Diodes should have same current and voltage rating as the SCR.

860 WATT LIMITED-RANGE LOW COST PRECISION LIGHT CONTROL


(860W)
430W
L1
L2
430W
(860W)
(240)
120V
60Hz
R1
D1
Q1
R2
C1
R3
Q2
D2
C2
R4
R5
P.C.
Z1

R_1 - 6.8KΩ, 2W (15K, 5W)
R_2 - 47Ω
R_3 - 1KΩ
R_4 - 680Ω
R_5 - 1KΩ, 1/2W POT
P.C. - GE A35

Q_1 - GE C122B(C122D)
Q_2 - GE 2N2646
D_1 - GE A41B (A41D)
D_2 - GE A14F
Z_1 - 16 V ZENER
L_1, L_2 - 430W INCANDESCENT LAMP

C_1 - .03μfd, 25V
C_2 - .01μfd, 25V
NOTE: ALL RESISTANCES 1/2w, 10% UNLESS OTHERWISE NOTED.
VALUES IN PARANTHESES FOR 240V

Fig. 45-11

Circuit Notes

The system is designed to regulate an 860 watt lamp load from half to full power. This is achieved by the controlled-half-plus-fixed-half-wave phase control method. Half power applied to an incandescent lamp results in 30% of the full light output. Consequently the circuit is designed to control the light output of the lamp from 30% to 100% of maximum.

800 W SOFT-START LIGHT DIMMER

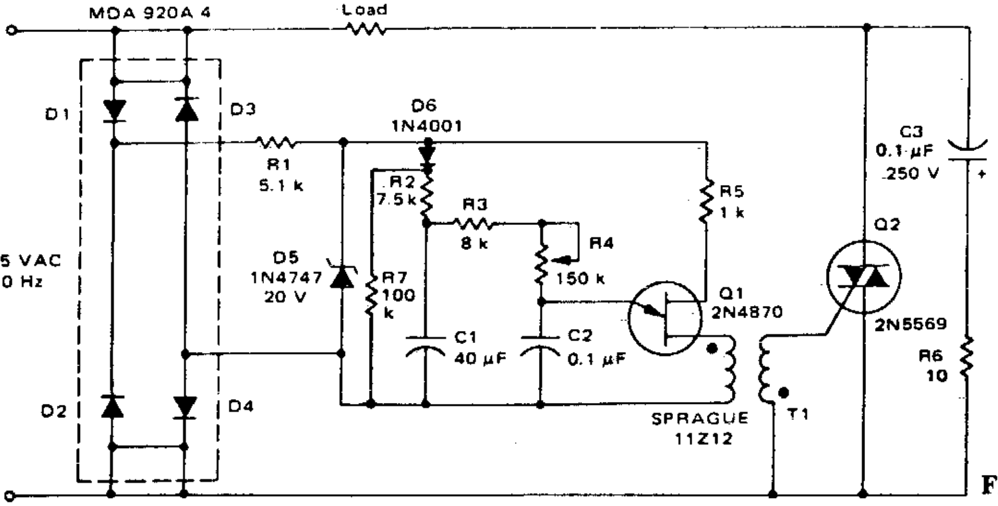
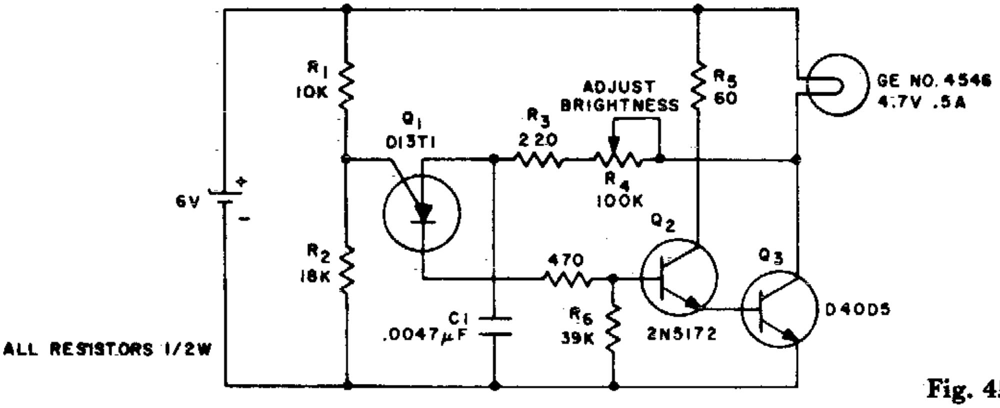

MDA 920A 4
Load
D1
D3
D2
D4
115 VAC
60 Hz
R1
5.1 k
D6
1N4001
D5
1N4747
20 V
R7
100
k
R2
7.5 k
R3
8 k
R4
150 k
C1
40 μF
C2
0.1 μF
R5
1 k
Q1
2N4870
SPRAGUE
11Z12
T1
C3
0.1 μF
250 V
+
Q2
2N5569
R6
10

Fig. 45-12

LOW LOSS BRIGHTNESS CONTROL



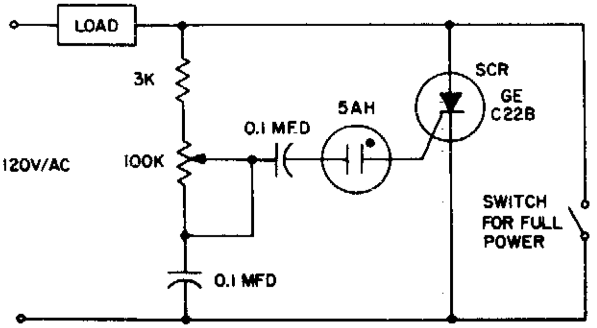
R1
10K
Q1
D13T1
R3
220
ADJUST
BRIGHTNESS
R5
60
GE NO. 4546
4.7V .5A
6V
+
-
R4
100K
Q2
R2
18K
470
Q3
C1
.0047μF
R6
39K
2N5172
D40D5
ALL RESISTORS 1/2W

Fig. 45-13

Circuit Notes

This circuit changes the average value of the dc supply voltage because of the high switching frequency. The tungsten lamp will have an almost continuous adjustable light output between 0 and 100%. If a light emitting diode is used as the emitting device, the irradiance will be in phase with the applied current pulses and will decrease to zero when the supply current is zero.

HALF WAVE AC PHASE-CONTROLLED CIRCUIT



LOAD
3K
SCR
GE
C22B
5AH
0.1 MFD
120V/AC
100K
SWITCH
FOR FULL
POWER
0.1 MFD

Fig. 45-14

Circuit Notes

The 5AH will trigger when the voltage across the two 0.1 μF capacitors reaches the breakdown voltage of the lamp. Control can be obtained full off to 95% of the half wave RMS output voltage. Full power can be obtained with the addition of the switch across the SCR.

EMERGENCY LIGHT

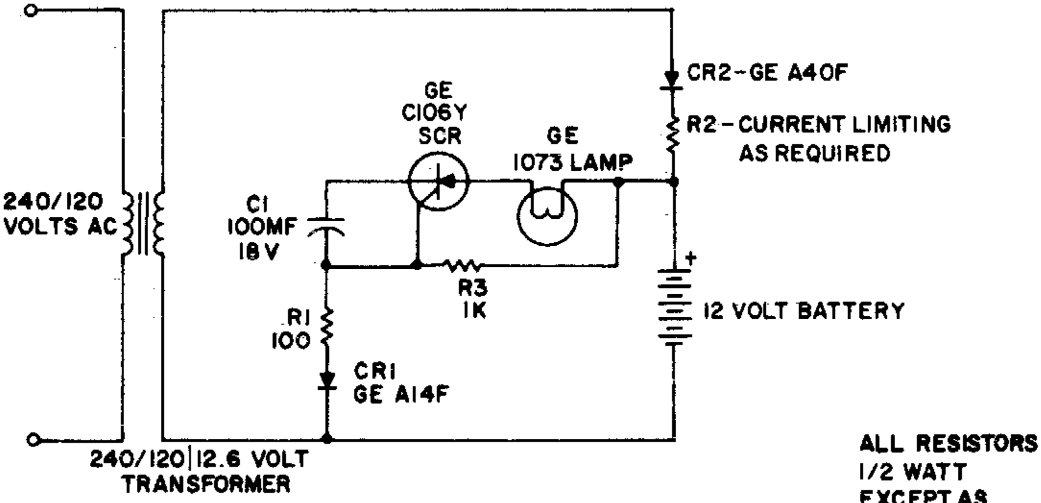

GE
C106Y
SCR
GE
1073 LAMP
CR2-GE A40F
R2-CURRENT LIMITING
AS REQUIRED
240/120
VOLTS AC
C1
100MF
18 V
R3
1K
+
12 VOLT BATTERY
R1
100
CR1
GE A14F
240/120|12.6 VOLT
TRANSFORMER
ALL RESISTORS
1/2 WATT
EXCEPT AS
NOTED

Fig. 45-15

Circuit Notes

This simple circuit provides battery operated emergency lighting instantaneously upon failure of the regular ac service. When line power is restored, the emergency light turns off and the battery recharges automatically. The circuit is ideal for use in elevator cars, corridors and similar places where loss of light due to power failure would be undesirable. Completely static in operation, the circuit requires no maintenance. With ac power on, capacitor C1 charges through rectifier CR1 and resistor R1 to develop a negative voltage at the gate of the C106Y SCR. By this means, the SCR is prevented from being triggered, and the emergency light stays off. At the same time, the battery is kept fully charged by rectifier CR2 and resistor R2. Should the ac power fail, C1 discharges and the SCR is triggered on by battery power through resistor R3. The SCR then energizes the emergency light. Reset is automatic when ac is restored, because the peak ac line voltage biases the SCR and turns it off.

NEON LAMP DRIVER

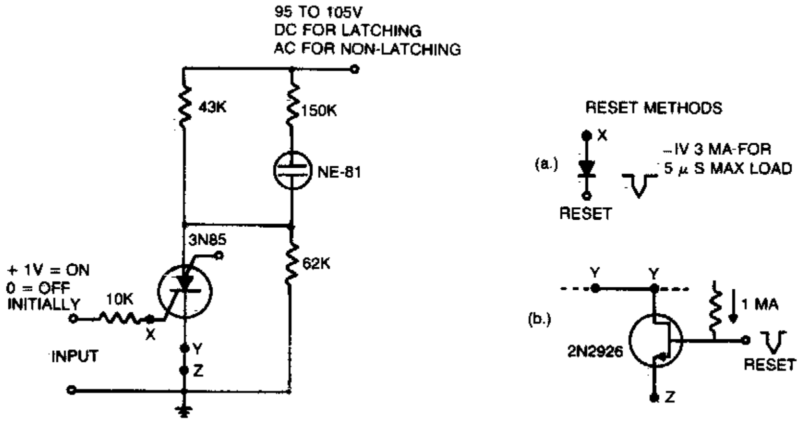

95 TO 105V
DC FOR LATCHING
AC FOR NON-LATCHING
43K
150K
NE-81
3N85
62K
+ 1V = ON
0 = OFF
INITIALLY
10K
X
INPUT
Y
Z
RESET METHODS
X
(a.)
RESET
-IV 3 MA FOR
5 μ S MAX LOAD
Y
Y
1 MA
(b.)
2N2926
RESET
Z

Fig. 45-16

COMPLEMENTARY AC POWER SWITCHING

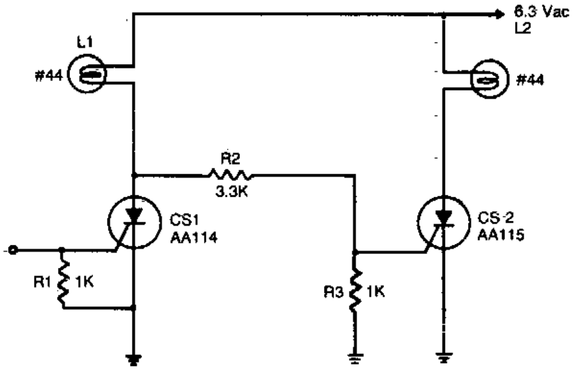

6.3 Vac
L2
L1
#44
#44
R2
3.3K
CS1
AA114
CS 2
AA115
R1
1K
R3
1K

Fig. 45-17

Circuit Notes

An input signal of less than 1 mA and 1 V is required to switch on CS1. As long as this input signal is maintained, CS1 will conduct during each positive half cycle of anode voltage, thereby energizing load L1 with half-wave rectified dc. L2 remains de-energized, since the anode of CS1 will not go more positive than 1.5 volts, and voltage divider R2 - R3 cannot provide enough voltage to trigger CS2. Upon removal of the input signal, CS1 will drop out. L1 will be de-energized, except for a small amount of ac current through R2 and R3. CS2 will be triggered on at the beginning of each positive half-cycle, when CS1 anode voltage reaches 2 to 3 volts. CS2 will conduct for nearly the entire positive half-cycle energizing L2. It should be noted that the 6.3 volt lamps used will operate at ⅓ the rated brilliance because of the controlled switch half-wave rectifying action and will extend the operating lamp life by several orders of magnitude. Should full brilliance be desired, the anode supply voltage level should be raised to 9 volts ac.

BATTERY LANTERN CIRCUIT

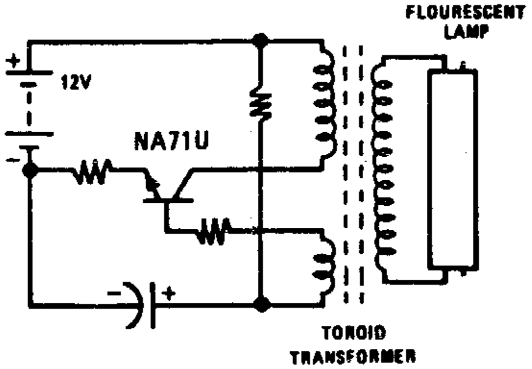

FLOURESCENT LAMP
+
12V
NA71U
TOROID TRANSFORMER

Fig. 45-18

SHIFT REGISTER

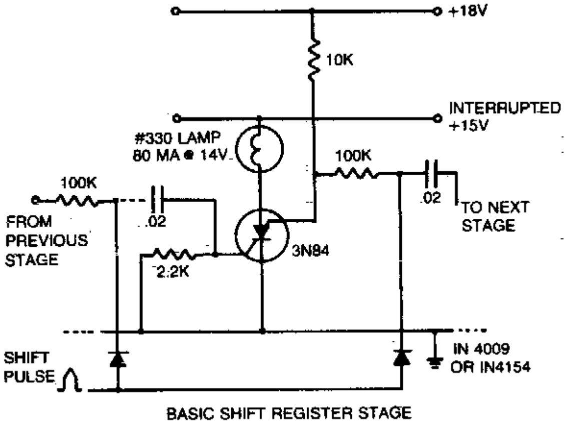

+18V
10K
INTERRUPTED +15V
#330 LAMP
80 MA @ 14V
100K
.02
TO NEXT STAGE
100K
FROM PREVIOUS STAGE
.02
2.2K
3N84
SHIFT PULSE
IN 4009 OR IN4154
BASIC SHIFT REGISTER STAGE

Fig. 45-19

Circuit Notes

The shift pulse amplitude is less than 15 volts. If a stage is off, the shift pulse will not be coupled to the next stage. If it is on, the diode will conduct and trigger the next stage. Just prior to the shift pulse the anode supply is interrupted to turn off all stages. The stored capacitor charge determines which stages will be triggered.

LIGHT-LEVEL CONTROLLER

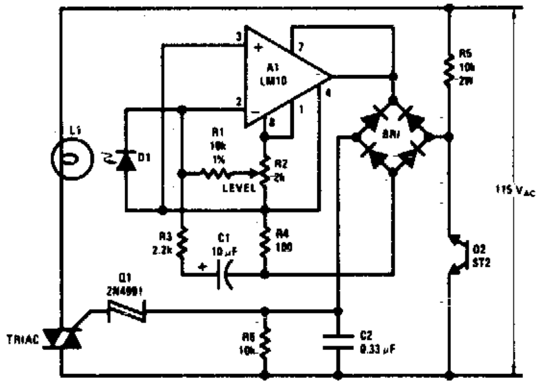

A1
LM10
3
2
7
4
8
1
L1
D1
R1
10k
1%
R2
2k
LEVEL
R3
2.2k
C1
10 μF
R4
100
BR1
R5
10k
2W
115 VAC
Q2
ST2
Q1
2N4991
TRIAC
R6
10k
C2
0.33 μF

Fig. 45-20

2.2 WATT INCANDESCENT LAMP DRIVER

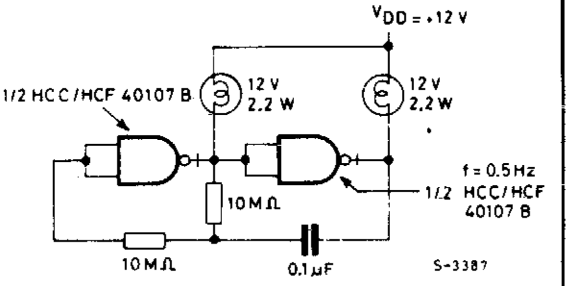

VDD = +12 V
1/2 HCC/HCF 40107 B
12 V
2.2 W
12 V
2.2 W
10 MΩ
f = 0.5 Hz
1/2 HCC/HCF 40107 B
10 MΩ
0.1 μF
S-3387

Fig. 45-21

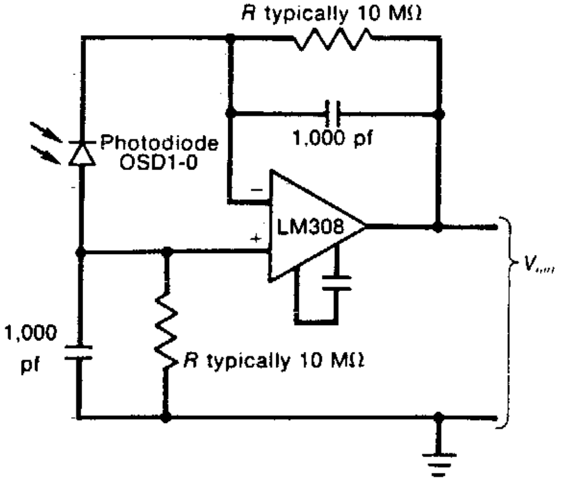
46

Light Measuring Circuits

The sources of the following circuits are contained in the Sources section beginning on page 730. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

- Linear Light Meter Circuit
- Logarithmic Light-Meter Circuit
- Light Meter
- Light Meter
- Light Meter
- Precision Photodiode Comparator

LINEAR LIGHT-METER CIRCUIT



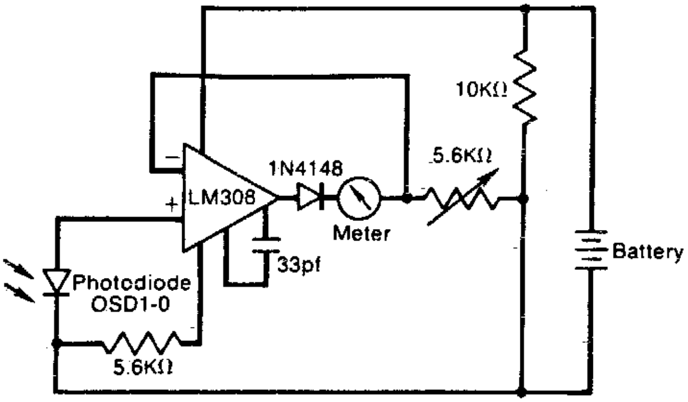
R typically 10 MΩ
1,000 pf
Photodiode
OSD1-0
LM308
1,000
pf
R typically 10 MΩ
V_{out}

Fig. 46-1

Circuit Notes

This circuit uses a low-input-bias op amp to give a steady dc indication of light level. To reduce circuit sensitivity to light, R1 can be reduced, but should not be less than 100 K. The capacitor values in the circuit are chosen to provide a time constant sufficient to filter high-frequency light variations that might arise, for example, from fluorescent lights.

LOGARITHMIC LIGHT-METER CIRCUIT



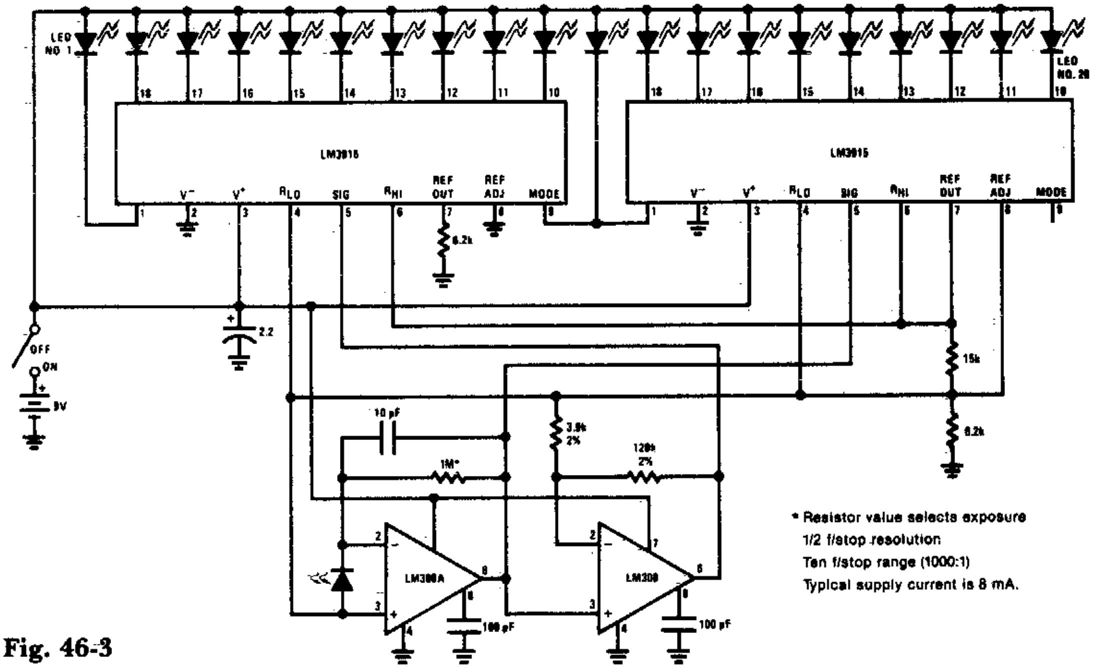
10KΩ
5.6KΩ
1N4148
LM308
Meter
33pf
Battery
Photodiode
OSD1-0
5.6KΩ

Fig. 46-2

Circuit Notes

The meter reading is directly proportional to the logarithm of the input light power. The logarithmic circuit behavior arises from the nonlinear diode pnjunction current/voltage relationship. The diode in the amplifier output prevents output voltage from becoming negative (thereby pegging the meter), which may happen at low lightlevels due to amplifier bias currents. R1 adjusts the meter full-scale deflection, enabling the meter to be calibrated.

LIGHT METER


LED NO 1
LED NO. 20
LM3915
LM3915
18
17
16
15
14
13
12
11
10
V⁻
V⁺
R LO
SIG
R HI
REF OUT
REF ADJ
MODE
1
2
3
4
5
6
7
8
9
8.2k
15k
8.2k
OFF
ON
9V
2.2
10 pF
1M*
3.9k 2%
120k 2%
LM308A
LM308
100 pF
100 pF

* Resistor value selects exposure
1/2 f/stop resolution
Ten f/stop range (1000:1)
Typical supply current is 8 mA.

Fig. 46-3

LIGHT METER

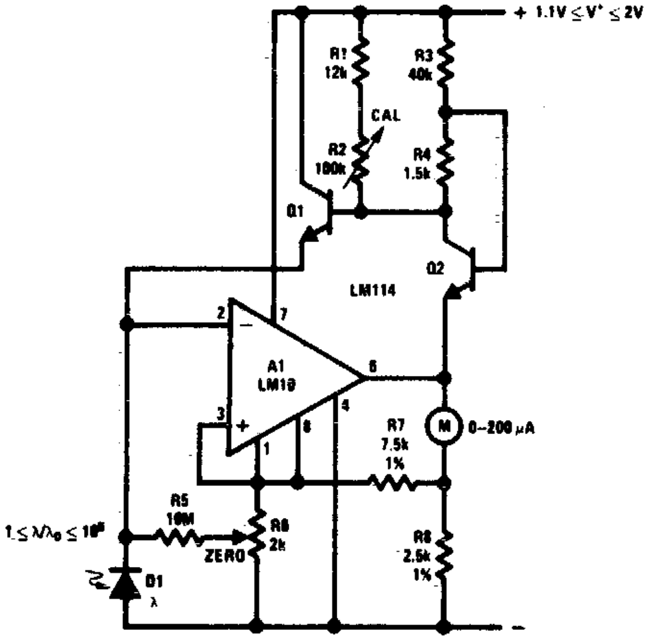
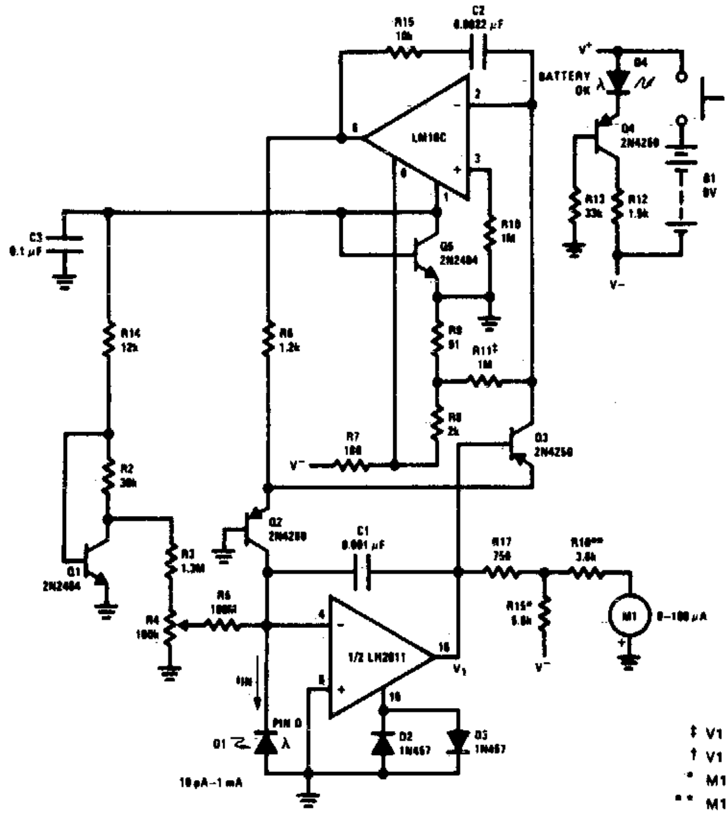

+ 1.1V ≤ V⁺ ≤ 2V
R1 12k
R3 40k
CAL
R2 100k
R4 1.5k
Q1
Q2
LM114
A1 LM10
0–200 μA
M
R7 7.5k 1%
R5 10M
R6 2k
ZERO
R8 2.5k 1%
1 ≤ λ/λ₀ ≤ 10⁵
D1 λ
2
3
7
6
4
8
1
−

Fig. 46-4

LIGHT METER


R15
10k
C2
0.0022 μF
V+
BATTERY
OK
D4
Q4
2N4250
B1
9V
R13
33k
R12
1.9k
V−
LM108
C3
0.1 μF
Q5
2N2484
R10
1M
R14
12k
R6
1.2k
R9
91
R11‡
1M
R8
2k
R7
100
V−
Q3
2N4250
R2
30k
Q2
2N4250
Q1
2N2484
R3
1.3M
C1
0.001 μF
R17
750
R16**
3.0k
R4
100k
R5
100M
R15*
5.6k
V−
M1
0–100 μA
1/2 LM2011
V1
IIN
PIN D
D1
D2
1N457
D3
1N457
10 pA–1 mA

‡ V1 = 0 @ I_{IN} = 100 nA

† V1 = −0.24V @ I_{IN} = 10 pA

* M1 = 0 @ I_{IN} = 10 pA

** M1 = I_{S} @ I_{IN} = 1 mA

Fig. 46-5

Circuit Notes

This light meter has an eight-decade range. Bias current compensation can give input current resolution of better than ±2 pA over 15 °C to 55 °C.

PRECISION PHOTODIODE COMPARATOR

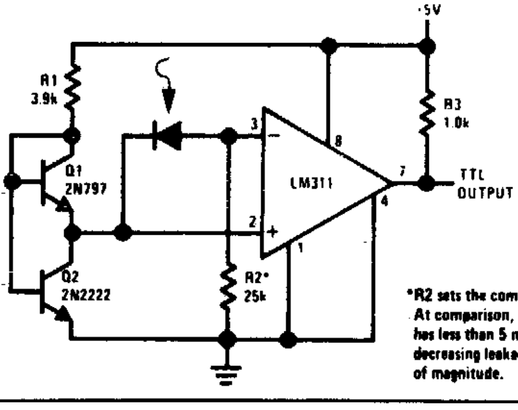

5V
R1
3.9k
Q1
2N797
Q2
2N2222
3
2
8
7
4
1
LM311
R2*
25k
R3
1.0k
TTL
OUTPUT

Fig. 46-6

*R2 sets the comparison level. At comparison, the photodiode has less than 5 mV across it, decreasing leakages by an order of magnitude.

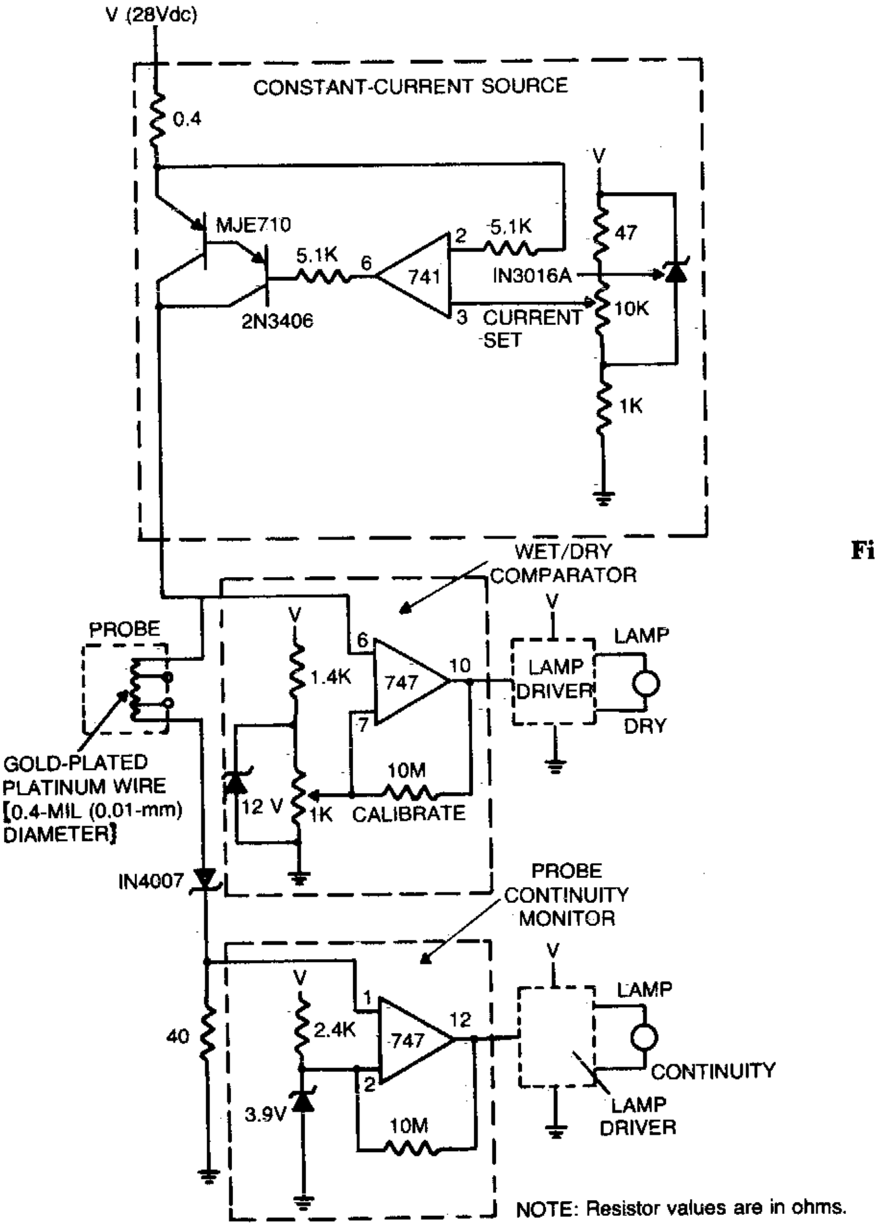
47

Liquid Level Detectors

The sources of the following circuits are contained in the Sources section beginning on page 730. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

- Level Sensor for Cryogenic Fluids
- Fluid Level Controller
- High Level Warning Device
- Liquid Level Control
- Liquid Level Detector Latching
- Water Level Alarm
- Water-Level Sensing Control Circuit
- Flood Alarm
- Liquid Level Detector
- Low-Level Warning with Audio Output

LEVEL SENSOR FOR CRYOGENIC FLUIDS


V (28Vdc)
CONSTANT-CURRENT SOURCE
0.4
V
MJE710
5.1K
6
741
2
5.1K
47
IN3016A
10K
3 CURRENT SET
2N3406
1K
WET/DRY COMPARATOR
V
PROBE
6
747
10
LAMP DRIVER
LAMP
DRY
1.4K
7
10M
GOLD-PLATED PLATINUM WIRE [0.4-MIL (0.01-mm) DIAMETER]
12 V
1K
CALIBRATE
IN4007
PROBE CONTINUITY MONITOR
V
V
LAMP
1
747
12
2.4K
40
2
CONTINUITY
3.9V
10M
LAMP DRIVER

NOTE: Resistor values are in ohms.

Fig. 47-1

Circuit Notes

The sensor circuit is adaptable to different liquids and sensors. The constant-current source drives current through the sensing probe and a fixed resistor. The voltage-comparator circuits interpret the voltage drops to tell whether the probe is immersed in liquid and whether there is current in the probe.

FLUID LEVEL CONTROLLER

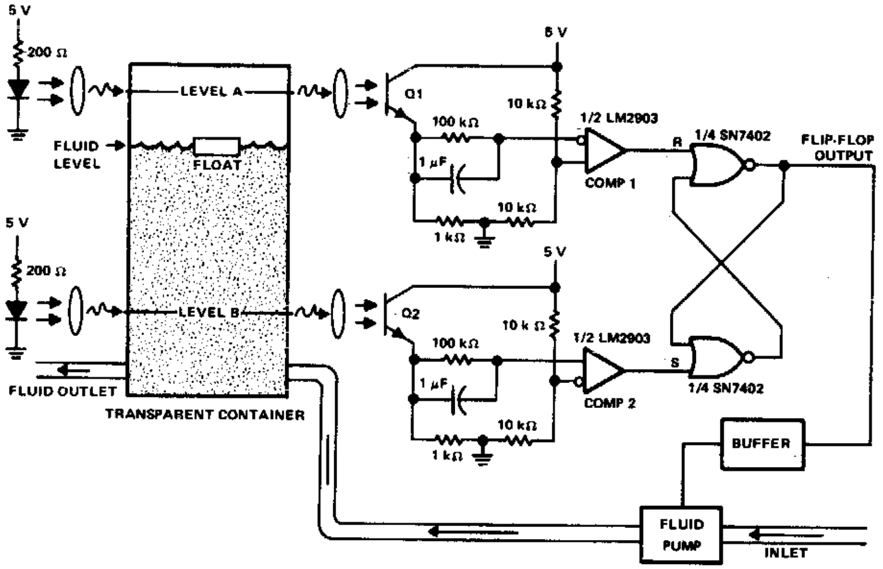
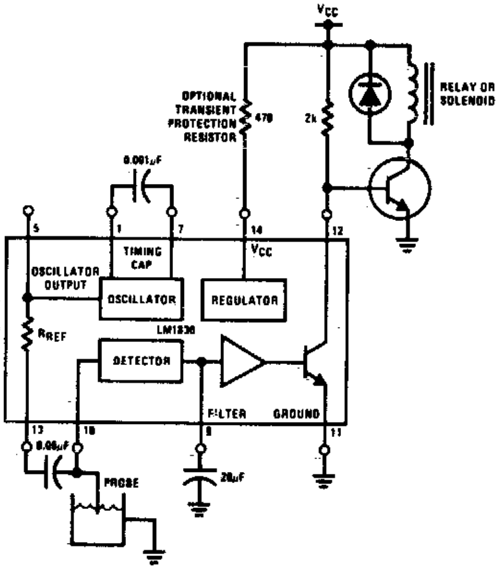

5 V
200 Ω
LEVEL A
FLUID LEVEL
FLOAT
5 V
200 Ω
LEVEL B
FLUID OUTLET
TRANSPARENT CONTAINER
5 V
Q1
10 kΩ
100 kΩ
1 µF
10 kΩ
1 kΩ
1/2 LM2903
COMP 1
R
1/4 SN7402
FLIP-FLOP OUTPUT
5 V
Q2
10 kΩ
100 kΩ
1 µF
10 kΩ
1 kΩ
1/2 LM2903
COMP 2
S
1/4 SN7402
BUFFER
FLUID PUMP
INLET

Fig. 47-2

Circuit Notes

This circuit can be used to maintain fluid between two levels. Variations on this control circuit can be made to keep something that moves within certain boundary conditions.

HIGH LEVEL WARNING DEVICE


V_{CC}
OPTIONAL TRANSIENT PROTECTION RESISTOR
470
2k
RELAY OR SOLENOID
0.001µF
5
1
7
14
12
TIMING CAP
V_{CC}
OSCILLATOR OUTPUT
OSCILLATOR
REGULATOR
LM1830
R_{REF}
DETECTOR
FILTER
GROUND
13
10
8
11
0.05µF
PROBE
20µF

The output is suitable for driving a sump pump or opening a drain valve, etc.

Fig. 47-3

LIQUID LEVEL CONTROL

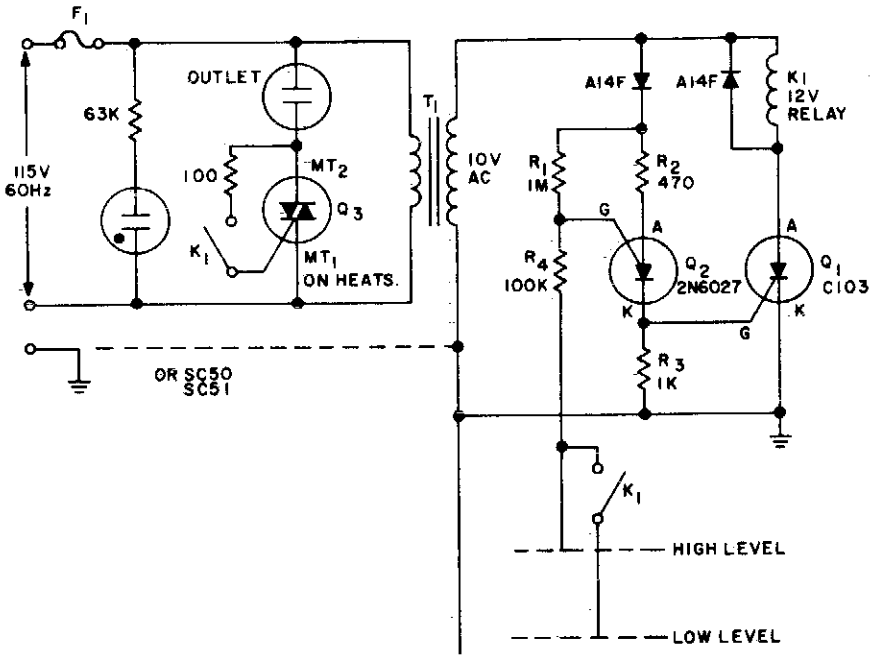
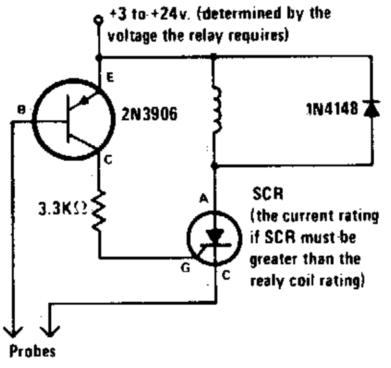

F1
115V 60Hz
63K
OUTLET
100
MT2
Q3
K1
MT1
ON HEATS.
T1
10V AC
OR SC50 SC51
A14F
A14F
K1 12V RELAY
R1 1M
R2 470
G
A
Q2 2N6027
Q1 C103
R4 100K
K
G
R3 1K
K1
HIGH LEVEL
LOW LEVEL

Fig. 47-4

Circuit Notes

Use this circuit to keep the fluid level of a liquid between two fixed points. Two modes, for filling or emptying are possible by simple reversing the contact connections of K1. The loads can be either electric motors or solenoid operated valves, operating from ac power. Liquid level detection is accomplished by two metal probes, one measuring the high level and the other the low level. An inversion of the logic (keeping the container filled) can be accomplished by replacing the normally open contact on the gate of Q3 with a normally closed contact.

LIQUID LEVEL DETECTOR (LATCHING)


+3 to +24v. (determined by the voltage the relay requires)
E
B
2N3906
C
1N4148
3.3KΩ
A
SCR (the current rating if SCR must be greater than the realy coil rating)
G
C
Probes

Circuit Notes

Alarm is actuated when liquid level is above the probes and remains activated even if the level drops below the probes. This latching action lets you know that the pre-set level has been reached or exceeded sometime in the past.

Fig. 47-5

WATER LEVEL ALARM

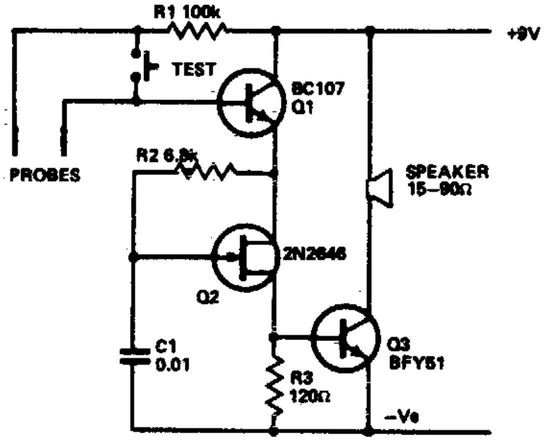

R1 100k
+9V
TEST
BC107
Q1
PROBES
R2 6.8k
SPEAKER
15–90Ω
2N2646
Q2
C1
0.01
R3
120Ω
Q3
BFY51
–Ve

Fig. 47-6

Circuit Notes

The circuit draws so little current that the shelf-line of the battery is the limiting factor. The only current drawn is the leakage of the transistor. The circuit is shown in the form of a water level alarm but by using different forms of probe can act as a rain alarm or shorting alarm; anything from zero to about 1 M between the probes will trigger it. Q1 acts as a switch which applies current to the unijunction relaxation oscillator Q2. Alarm signal frequency is controlled by values and ratios of C1/R2. Pulses switch Q3 on and off, applying a signal to the speaker. Almost any NPN silicon transistor can be used for Q1 and Q3 and almost any unijunction for Q2.

WATER-LEVEL SENSING CONTROL CIRCUIT

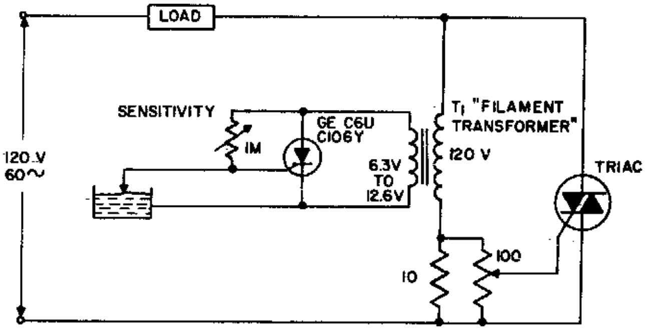
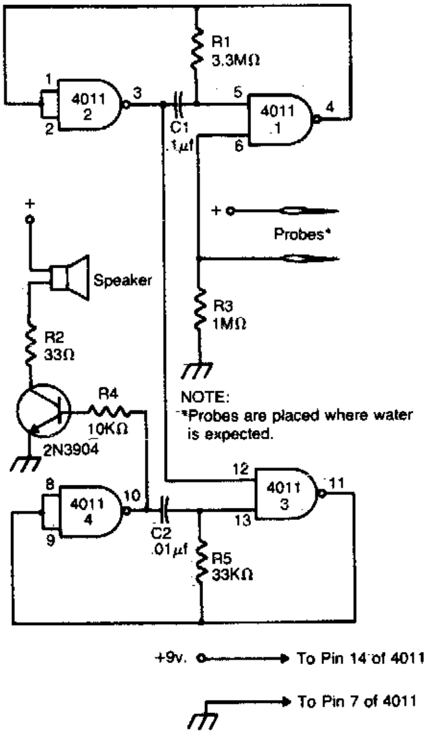

LOAD
SENSITIVITY
1M
GE C6U
C106Y
6.3V
TO
12.6V
T1 "FILAMENT
TRANSFORMER"
120 V
120 V
60~
TRIAC
100
10

Fig. 47-7

Circuit Notes

The circuit applies power to the load until the water conducts through the probe, and bypasses gate current from the low current SCR. This gives an isolated low voltage probe to satisfy safety requirements.

FLOOD ALARM


R1
3.3MΩ
1
4011
2
3
5
4011
1
4
2
C1
.1µf
6
+
+
Probes*
Speaker
R3
1MΩ
R2
33Ω
R4
10KΩ
2N3904
12
4011
3
11
8
4011
4
10
13
9
C2
.01µf
R5
33KΩ
+9v.
To Pin 14 of 4011
To Pin 7 of 4011

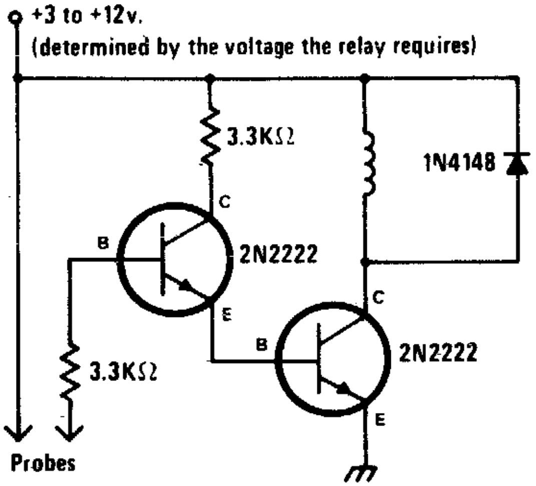
NOTE:
*Probes are placed where water is expected.

Circuit Notes

The alarm is built around two audio oscillators, each using two NAND gates. The detection oscillator is gated on by a pair of remote probes. One of the probes is connected to the battery supply, the other to the input of one of the gates. When water flows between the probes, the detection oscillator is gated on. The alarm oscillator is gated on by the output of the detection oscillator. The values given produce an audio tone of about 3000 Hz. The detection oscillator gates this audio tone at a rate of about 3 Hz. The result is a unique pulsating note. Use any 8 ohm speaker to sound the alarm. The 2N3904 can be replaced by any similar NPN transistor. The circuit will work from any six to 12-volt supply.

Fig. 47-8

LIQUID LEVEL DETECTOR

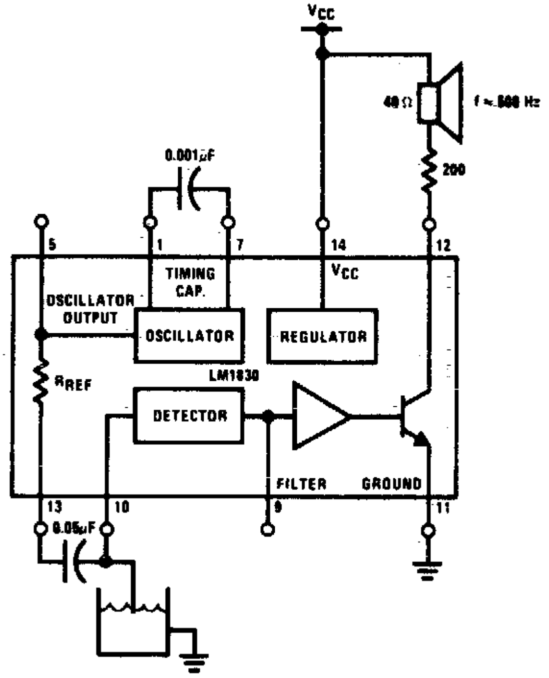

+3 to +12v.
(determined by the voltage the relay requires)
3.3KΩ
1N4148
C
B
2N2222
E
C
B
2N2222
E
3.3KΩ
Probes

Circuit Notes

When liquid level reaches both probes, alarm is turned on. When water level recedes it goes off.

Fig. 47-9

LOW-LEVEL WARNING WITH AUDIO OUTPUT



V_{CC}
40 Ω
f ≈ 600 Hz
200
0.001μF
5
1
7
14
12
TIMING CAP.
V_{CC}
OSCILLATOR OUTPUT
OSCILLATOR
REGULATOR
LM1830
R_{REF}
DETECTOR
FILTER
GROUND
13
10
9
11
0.05μF

Fig. 47-10

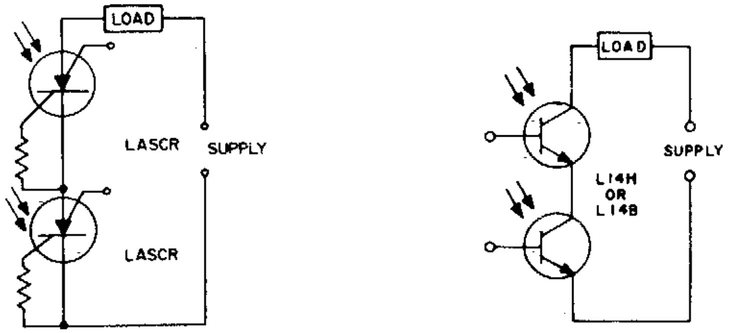
48

Logic Circuits

The sources of the following circuits are contained in the Sources section beginning on page 730. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

Light Activated Logic Circuits
Programmable Gate
Negative to Positive Supply Logic Level Shifter
OR Gate
OR Gate
Large Fan-In AND Gate
AND Gate
R-S Flip-Flop
AND Gate

LIGHT ACTIVATED LOGIC CIRCUITS


LOAD
LASCR
SUPPLY
LASCR
LOAD
SUPPLY
L14H
OR
L14B

(a) AND Circuit **(b) AND Circuit**

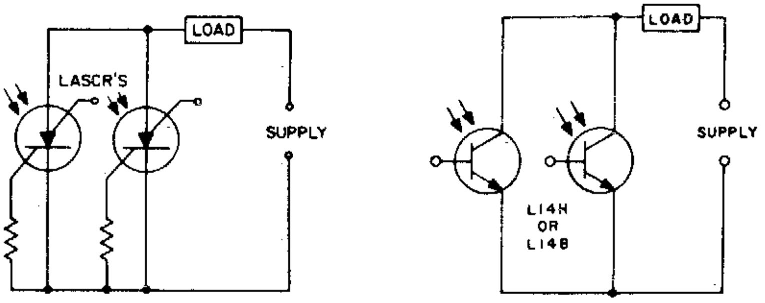
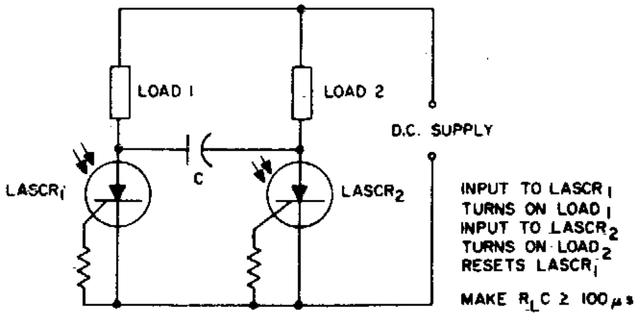

LOAD
LASCR'S
SUPPLY
LOAD
SUPPLY
L14H
OR
L14B

Fig. 48-1

(c) OR Circuit **(d) OR Circuit**


LOAD 1
LOAD 2
D.C. SUPPLY
C
$LASCR_1$
$LASCR_2$
INPUT TO $LASCR_1$
TURNS ON $LOAD_1$
INPUT TO $LASCR_2$
TURNS ON $LOAD_2$
RESETS $LASCR_1$
MAKE $R_LC \geq 100\mu s$

(e) Flip-Flop

Circuit Notes

These circuits illustrate some of the common logic functions that can be implemented.

PROGRAMMABLE GATE

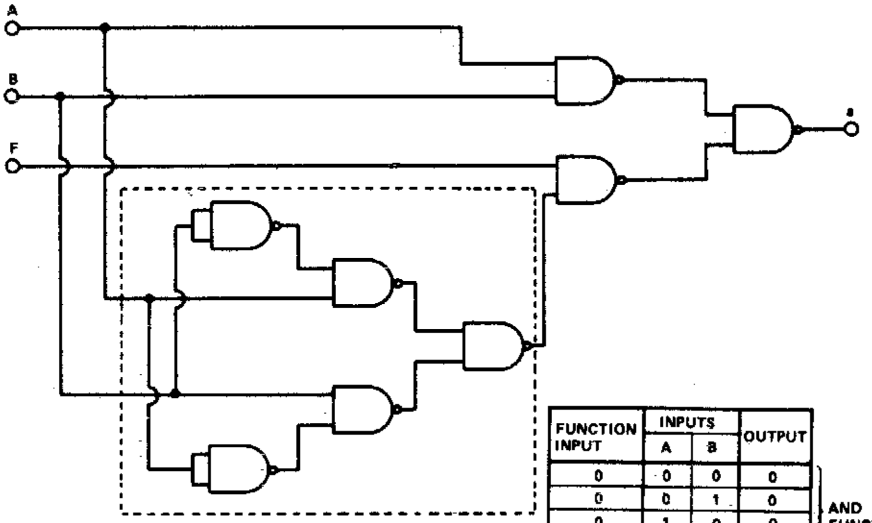


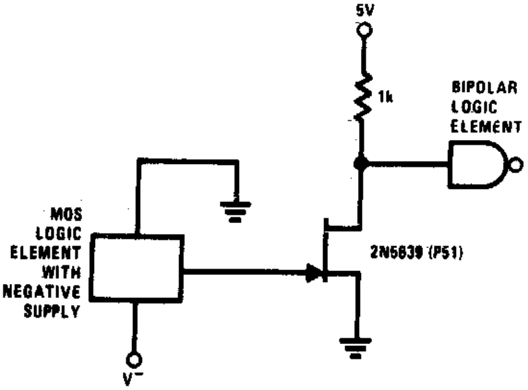
Fig. 48-2

FUNCTION INPUT	INPUTS A	INPUTS B	OUTPUT	
0	0	0	0	AND FUNCTION
0	0	1	0	
0	1	0	0	
0	1	1	1	
1	0	0	0	OR FUNCTION
1	0	1	1	
1	1	0	1	
1	1	1	1	

Circuit Notes

This gate converts an AND gate or an OR gate by applying a logic '1' on the function input. The logic design uses 8 two-input NAND gates. The number of gates may be reduced by replacing the 5 NAND gates enclosed by the dotted line with a two-input exclusive-OR, such as the TTL 7486.

NEGATIVE TO POSITIVE SUPPLY LOGIC LEVEL SHIFTER


5V
1k
BIPOLAR LOGIC ELEMENT
MOS LOGIC ELEMENT WITH NEGATIVE SUPPLY
2N5639 (P51)
V⁻

Circuit Notes

This simple circuit provides for level shifting from any logic function (such as MOS) operating from minus to ground supply to any logic level (such as TTL) operating from a plus to ground supply. The 2N5639 provides a low $r_{dc\,(ON)}$ and fast switching times.

Fig. 48-3

OR GATE

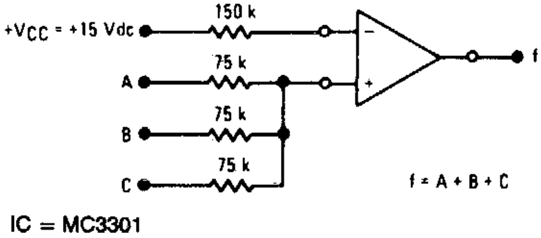

+VCC = +15 Vdc
150 k
75 k
75 k
75 k
A
B
C
f
f = A + B + C
IC = MC3301

Fig. 48-4

OR GATE

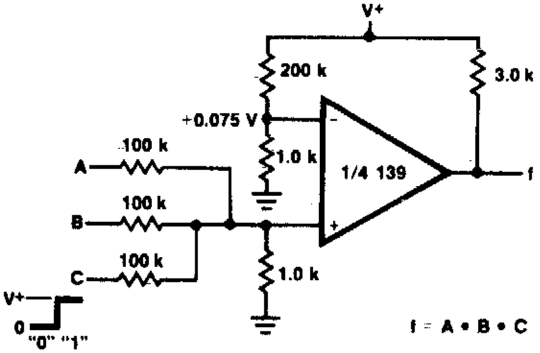

V+
200 k
3.0 k
+0.075 V
1.0 k
1/4 139
100 k
100 k
100 k
A
B
C
f
1.0 k
V+
0
"0" "1"
f = A • B • C

Fig. 48-5

LARGE FAN-IN AND GATE

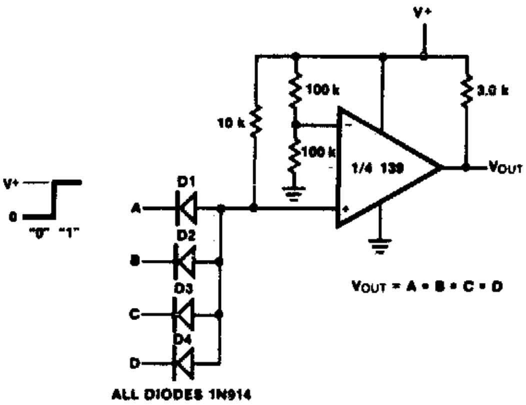

V+
10 k
100 k
100 k
3.0 k
1/4 139
VOUT
V+
0
"0" "1"
D1
D2
D3
D4
A
B
C
D
VOUT = A • B • C • D
ALL DIODES 1N914

Fig. 48-6

AND GATE

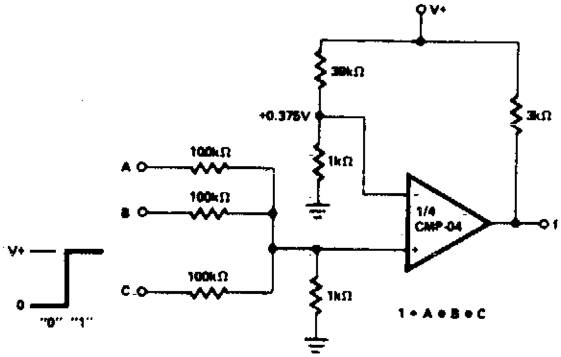

V+
39kΩ
3kΩ
+0.375V
1kΩ
100kΩ
100kΩ
100kΩ
A
B
C
1/4 CMP-04
f
1kΩ
V+
0
"0" "1"
f = A • B • C

Fig. 48-7

R-S FLIP-FLOP

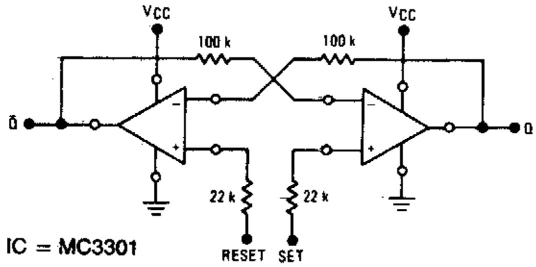

VCC
VCC
100 k
100 k
Q̄
Q
22 k
22 k
RESET
SET
IC = MC3301

Fig. 48-8

AND GATE

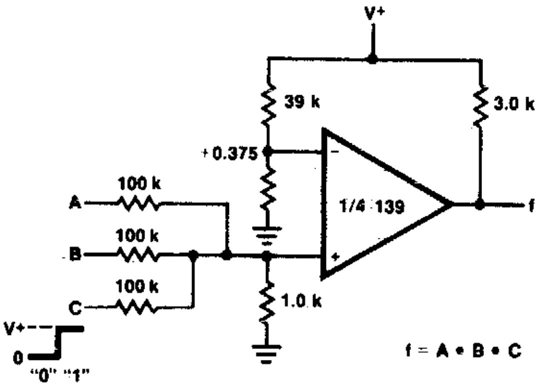

V+
39 k
3.0 k
+0.375
1/4 139
100 k
100 k
100 k
A
B
C
f
1.0 k
V+
0
"0" "1"
f = A • B • C

Fig. 48-9

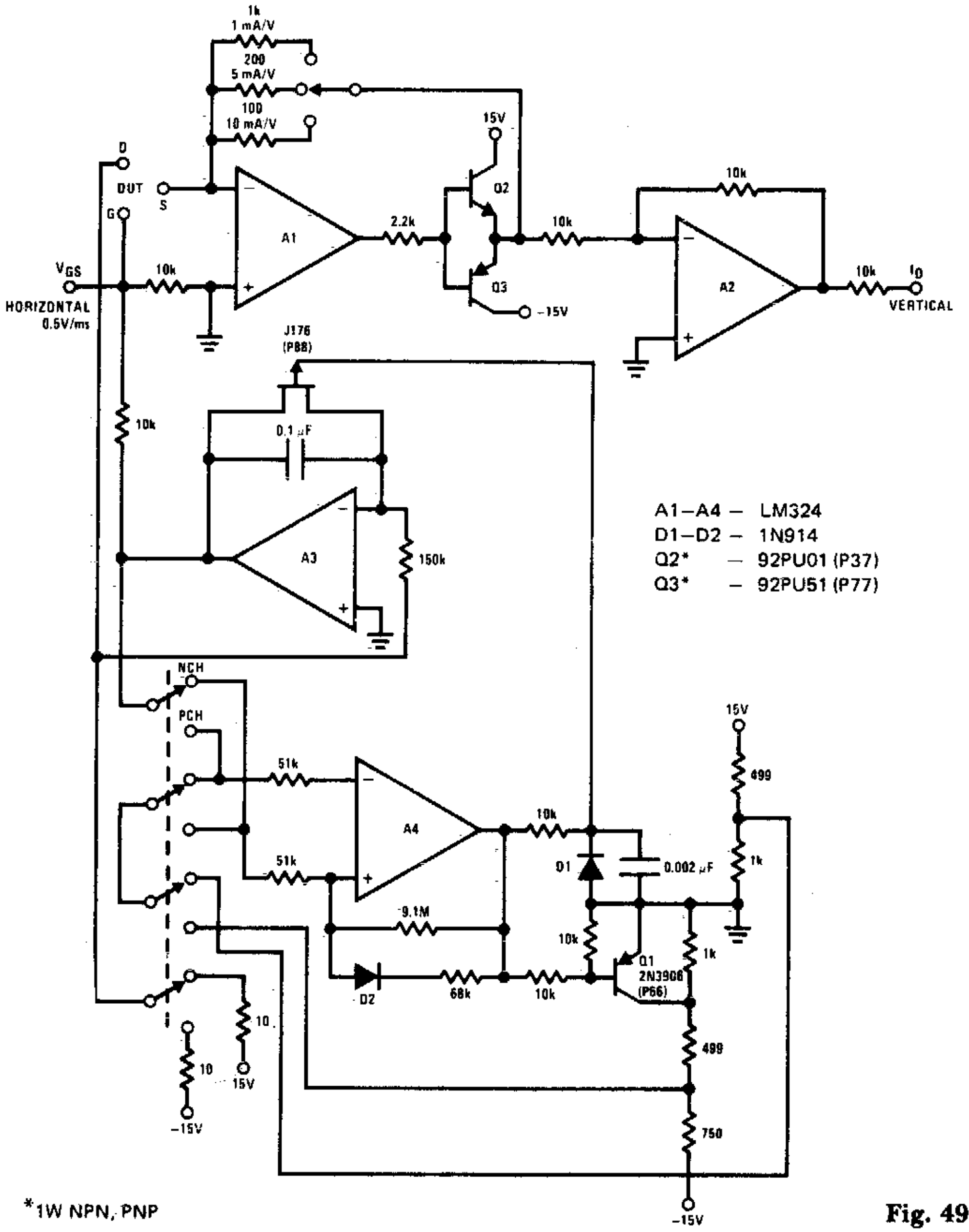
49

Measuring Circuits

The sources of the following circuits are contained in the Sources section beginning on page 730. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

FET Curve Tracer
Digital Weight Scale
Low Cost pH Meter
pH Probe Amplifier/Temperature Compensator
Capacitance Meter
Zener Tester
Transistor Sorter/Tester
Go/No-Go Diode Tester
Diode Tester
Peak Level Indicator
Sound Level Monitor
Linear Variable Differential Transformer (LVDT) Driver Demodulator
Linear Variable Differential Transformer (LVDT) Measuring Gauge
Vibration Meter
Sensitive RF Voltmeter
Minimum Component Tachometer
Phase Meter
Precision Calibration Standard
Zener Diode Checker

FET CURVE TRACER


1k
1 mA/V
200
5 mA/V
100
10 mA/V
15V
D
DUT
S
G
Q2
A1
2.2k
10k
10k
A2
VGS
HORIZONTAL
0.5V/ms
10k
Q3
-15V
10k
ID
VERTICAL
J176
(P88)
10k
0.1 μF
A3
150k
NCH
PCH
51k
A4
10k
15V
499
51k
D1
0.002 μF
1k
9.1M
10k
Q1
2N3906
(P66)
1k
D2
68k
10k
499
10
10
15V
750
-15V
-15V

A1–A4 – LM324
D1–D2 – 1N914
Q2* – 92PU01 (P37)
Q3* – 92PU51 (P77)

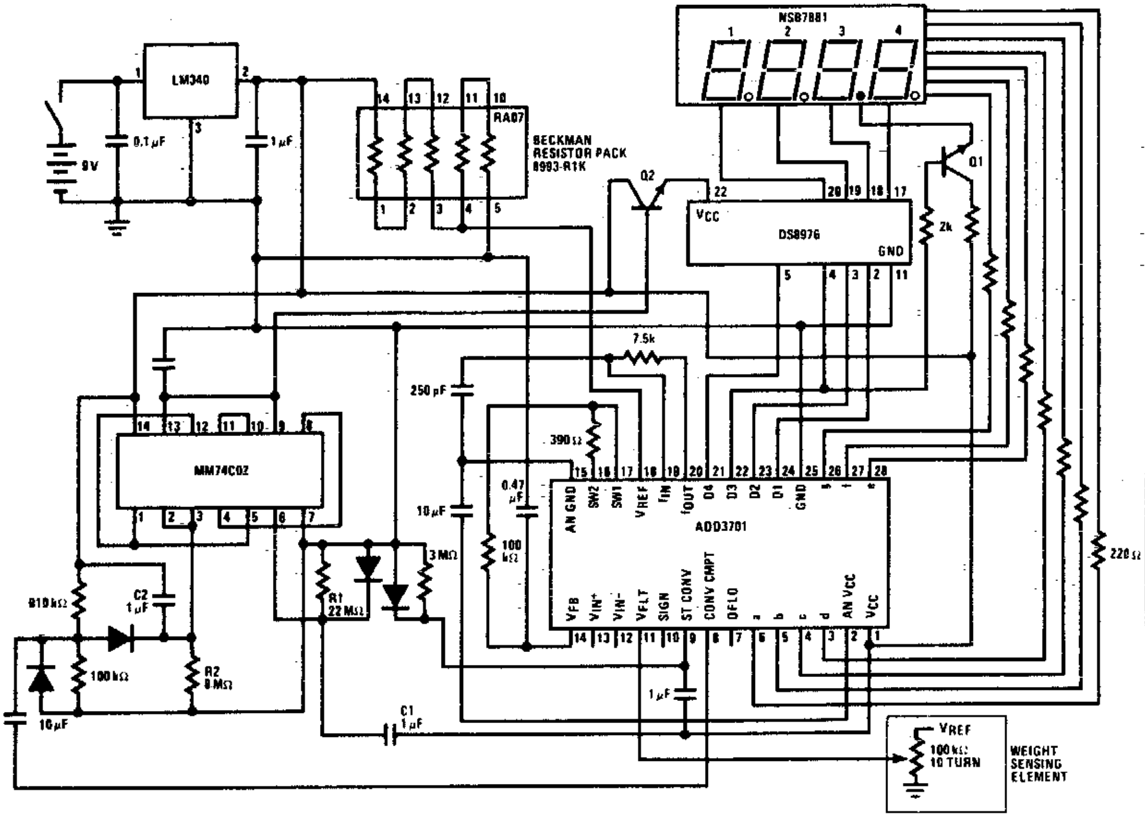
*1W NPN, PNP

Fig. 49-1

Circuit Notes

The circuit displays drain current versus gate voltage for both P and N-channel JFETS at a constant drain voltge.

DIGITAL WEIGHT SCALE


NSB7881
1
2
3
4
LM340
1
2
3
0.1µF
1µF
9V
14
13
12
11
10
RA07
BECKMAN
RESISTOR PACK
8993-R1K
1
2
3
4
5
Q2
22
VCC
DS8976
GND
20
19
18
17
5
4
3
2
11
Q1
2k
7.5k
250pF
390Ω
MM74C02
14
13
12
11
10
9
8
1
2
3
4
5
6
7
0.47 µF
10µF
100 kΩ
3MΩ
15 16 17 18 19 20 21 22 23 24 25 26 27 28
AN GND
SW2
SW1
VREF
IIN
IOUT
D4
D3
D2
D1
GND
g
f
e
ADD3701
VFB
VIN+
VIN−
VFLT
SIGN
ST CONV
CONV CMPT
OFLO
a
b
c
d
AN VCC
VCC
14 13 12 11 10 9 8 7 6 5 4 3 2 1
810kΩ
C2
1µF
100kΩ
R2
8MΩ
R1
22MΩ
10µF
C1
1µF
1µF
VREF
100kΩ
10 TURN
WEIGHT
SENSING
ELEMENT
220Ω

Notes:

1. R1, C1 defines POWER ON display blanking interval. R2, C2 defines display ON time.
2. All V_{CC} connections should use a single V_{CC} point and all ground/analog ground connections should use a single ground/analog ground point.
3. Display sequence for Rev A ckt implementation:

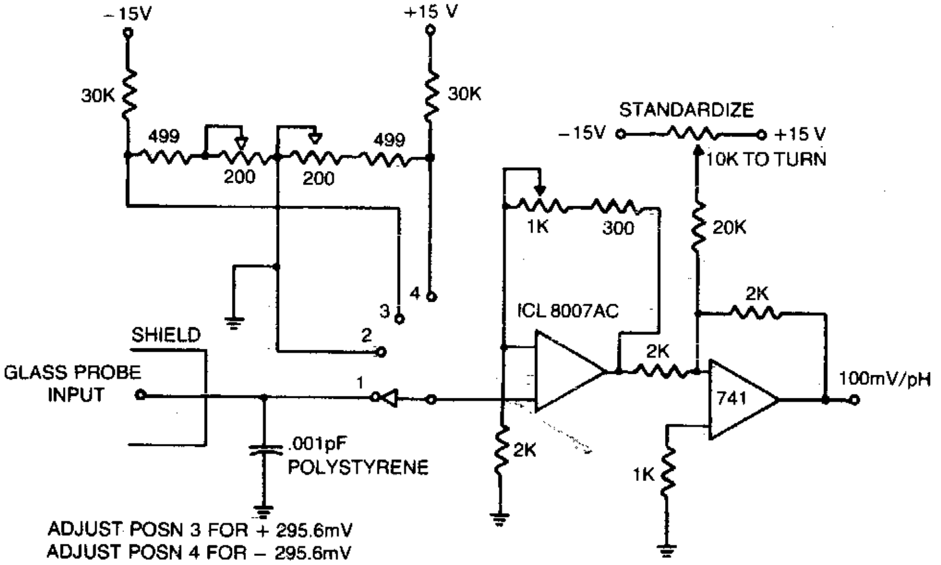
t = 0 sec • power ON
t = 0 → 5 sec • display blanked
• system converging
t = 5 → 10 sec • conversion complete
• display ENABLE
t ⩾ 10 sec • display blanked
• wait for new POWER UP cycle

Fig. 49-2

Circuit Notes

This circuit employs a potentiometer as the weight sensing element. An object placed upon the scale displaces the potentiometer wiper, an amount proportional to its weight. Conversion of the wiper voltage to digital information is performed, decoded, and interfaced to the numeric display.

LOW COST pH METER



−15V
+15 V
30K
30K
499
499
200
200
STANDARDIZE
−15V
+15 V
10K TO TURN
1K
300
20K
2K
ICL 8007AC
SHIELD
GLASS PROBE INPUT
1
2
3
4
2K
741
100mV/pH
.001pF POLYSTYRENE
2K
1K

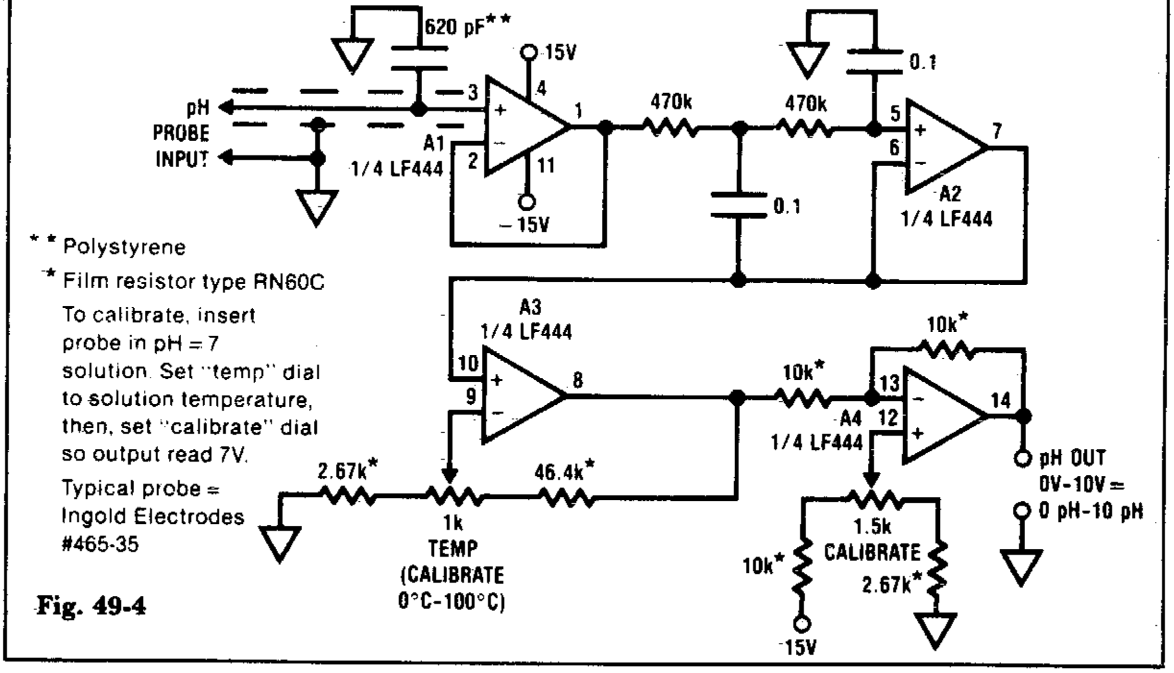
ADJUST POSN 3 FOR + 295.6mV
ADJUST POSN 4 FOR − 295.6mV

Fig. 49-3

Circuit Notes

With guaranteed 1 pA input bias, the ICL 8007A is ideal as a pH meter or long term sample and hold.

pH PROBE AMPLIFIER/TEMPERATURE COMPENSATOR



620 pF**
15V
pH PROBE INPUT
3
4
1
2
11
A1
1/4 LF444
−15V
470k
470k
0.1
0.1
5
6
7
A2
1/4 LF444
A3
1/4 LF444
10
9
8
10k*
10k*
13
12
14
A4
1/4 LF444
pH OUT
0V-10V = 0 pH-10 pH
2.67k*
46.4k*
1k
TEMP (CALIBRATE 0°C-100°C)
1.5k
CALIBRATE
10k*
2.67k*
−15V

** Polystyrene

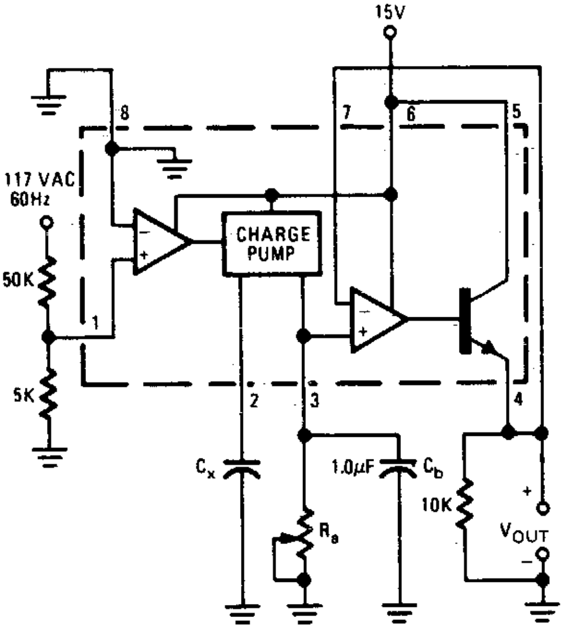
* Film resistor type RN60C

To calibrate, insert probe in pH = 7 solution. Set "temp" dial to solution temperature, then, set "calibrate" dial so output read 7V.

Typical probe = Ingold Electrodes #465-35

Fig. 49-4

CAPACITANCE METER


15V
8
7
6
5
117 VAC
60Hz
50K
5K
1
CHARGE PUMP
2
3
4
C_x
1.0μF
C_b
10K
R_a
+
V_{OUT}
−

Circuit Notes

Output voltage is proportional to the capacitance connected to pin 2 of the charge pump. The meter works over a range of 0.01 to 0.1 μF with R_a set at 111 K. Over this range of capacitance, the output voltage varies from 1 to 10 volts with a 15 volt power supply. A constant frequency reference is taken from the 60-Hz line.

Fig. 49-5

ZENER TESTER

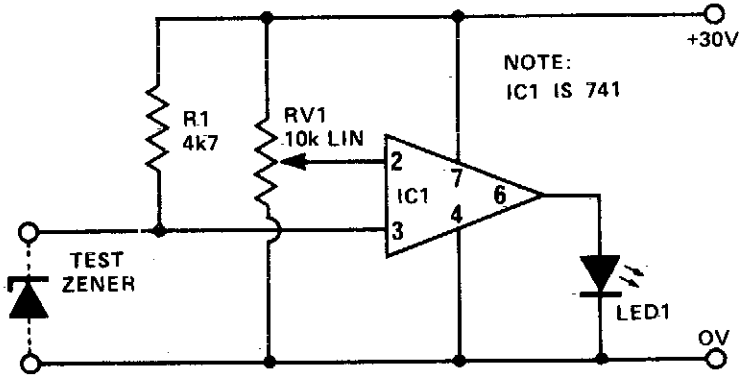

+30V
NOTE:
IC1 IS 741
R1
4k7
RV1
10k LIN
2
3
7
6
4
IC1
TEST ZENER
LED1
0V

Fig. 49-6

Circuit Notes

This circuit provides a low cost and reliable method of testing zener diodes. RV1 can be calibrated in volts, so that when LED 1 just lights, the voltage on pins 2 and 3 are nearly equal. Hence, the zener voltage can be read directly from the setting of RV1. The supply need only be as high a value as the zener itself. For a more accurate measurement, a precision pot could be added and calibrated.

TRANSISTOR SORTER/TESTER

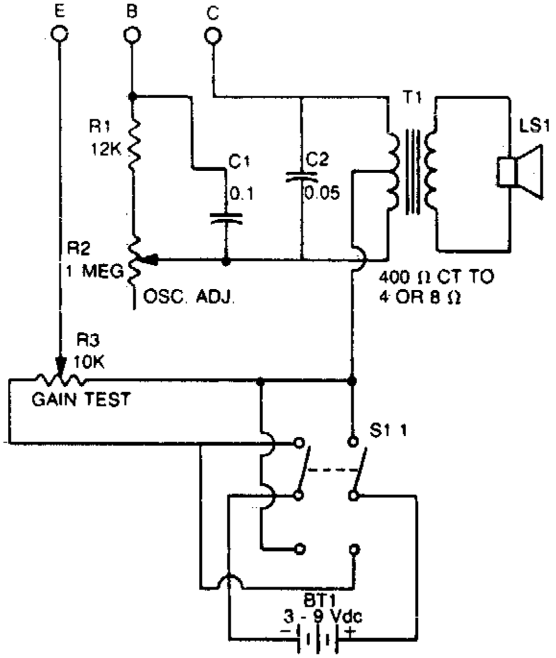

E
B
C
T1
LS1
R1
12K
C1
0.1
C2
0.05
R2
1 MEG
OSC. ADJ.
400 Ω CT TO
4 OR 8 Ω
R3
10K
GAIN TEST
S1 1
BT1
3 - 9 Vdc
-
+

Fig. 49-7

Circuit Notes

This tester checks transistor for polarity (PNP or NPN). An audible signal will give an indication of gain. Tester can also be used as a GO/NO GO tester to match unmarked devices.

GO/NO-GO DIODE TESTER

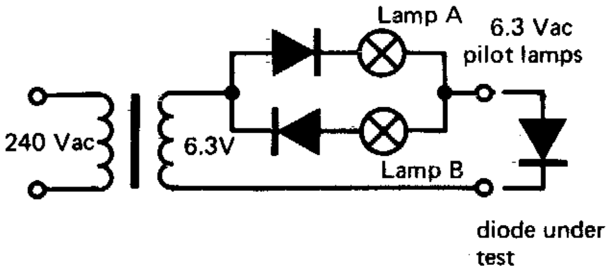
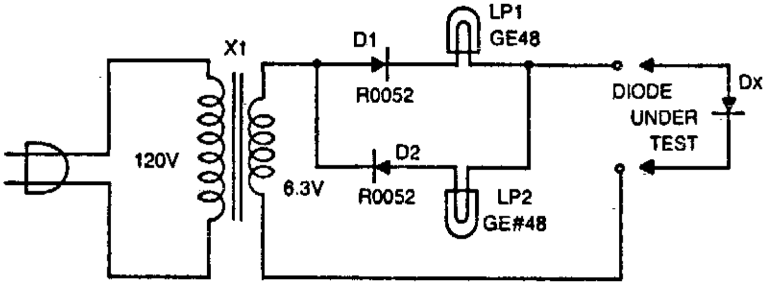

Lamp A
6.3 Vac
pilot lamps
240 Vac
6.3V
Lamp B
diode under
test

Fig. 49-8

Circuit Notes

If lamp A or B is illuminated, the diode is serviceable. If both light, the diode is short circuited. If neither light, diode is an open circuit.

DIODE TESTER


X1
120V
6.3V
D1
R0052
D2
R0052
LP1
GE48
LP2
GE#48
DIODE
UNDER
TEST
DX

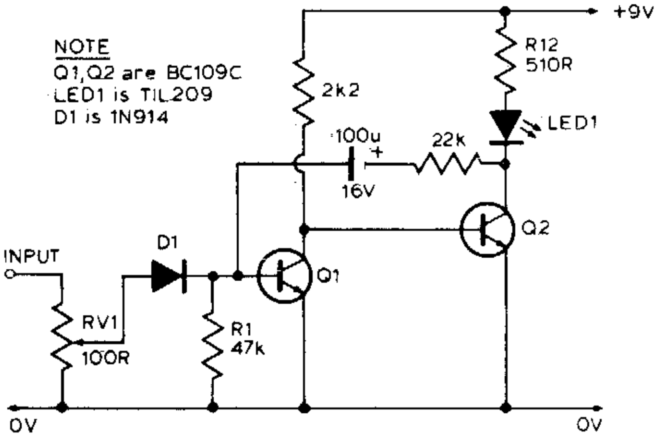
Parts list
D1, D2—HEP R0052
LP1, LP2—GE 48 lamp
X1—120V to 6.3V, transformer, Lafayette 33P80508

Fig. 49-9

Circuit Notes

The circuit tests whether or not a diode is open, shorted, or functioning correctly. If lamp A lights, the diode under test is functional. When lamp B is lit, the diode is good but connected backwards. When both lamps are lit, the diode is shorted, and it is open if neither lamp is lit.

PEAK LEVEL INDICATOR


+9V
R12
510R
2k2
LED1
100u
16V
22k
Q2
Q1
INPUT
D1
RV1
100R
R1
47k
0V
0V

NOTE
Q1,Q2 are BC109C
LED1 is TIL209
D1 is 1N914

Fig. 49-10

Circuit Notes

The LED is normally lit, but it will be briefly extinguished if the input exceeds a preset (by RV1) level. A possible application is to monitor the output voltage across a loudspeaker; the LED will flicker with large signals.

SOUND LEVEL MONITOR

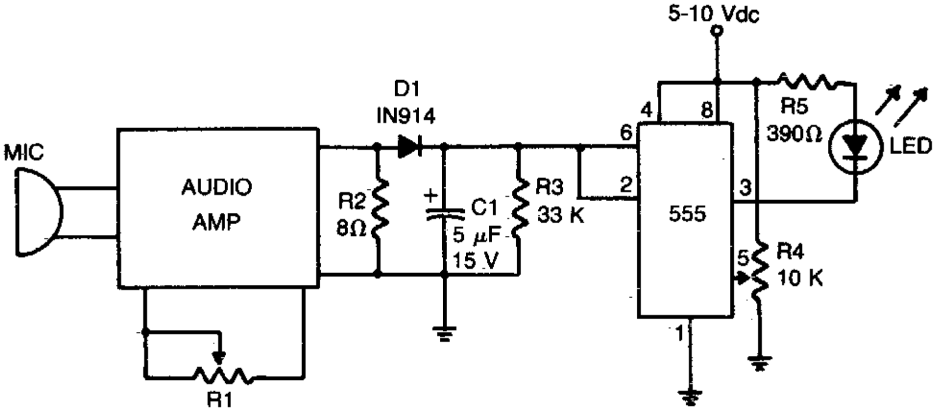

5-10 Vdc
D1
IN914
MIC
AUDIO
AMP
R2
8Ω
+
C1
5 μF
15 V
R3
33 K
4
8
6
2
3
5
1
555
R5
390Ω
LED
R4
10 K
R1

Fig. 49-11

Circuit Notes

Loudness detector consists of a 555 IC wired as a Schmitt trigger. The output changes state—from high to low—whenever the input crosses a certain voltage. That threshold voltage is established by the setting of R4.

LINEAR VARIABLE DIFFFERENTIAL TRANSFORMER (LVDT) DRIVER DEMODULATOR

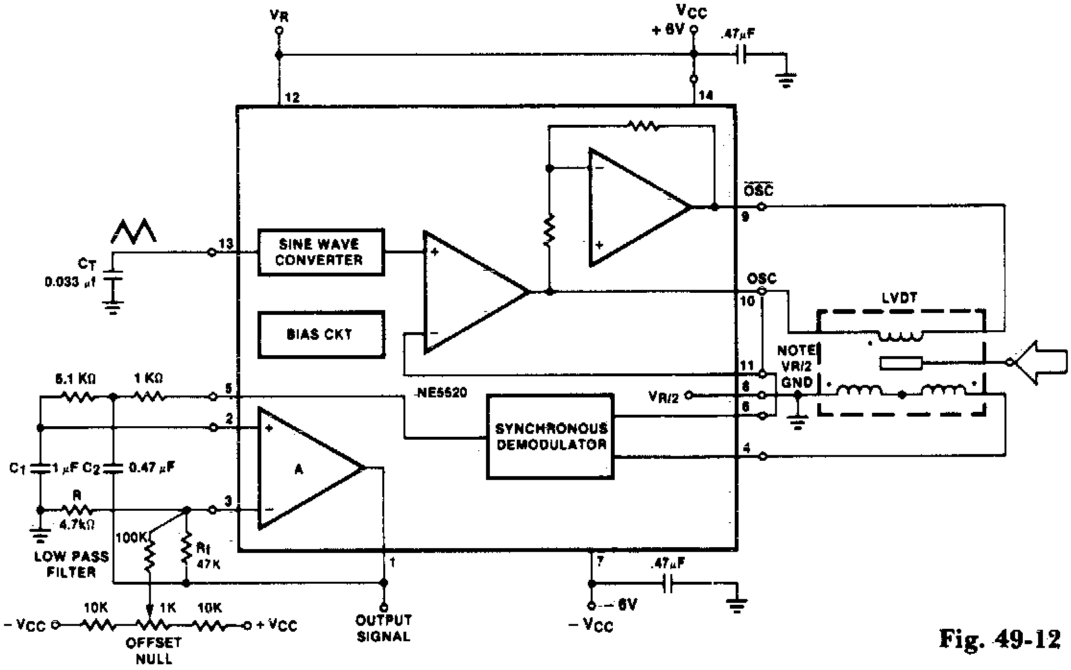
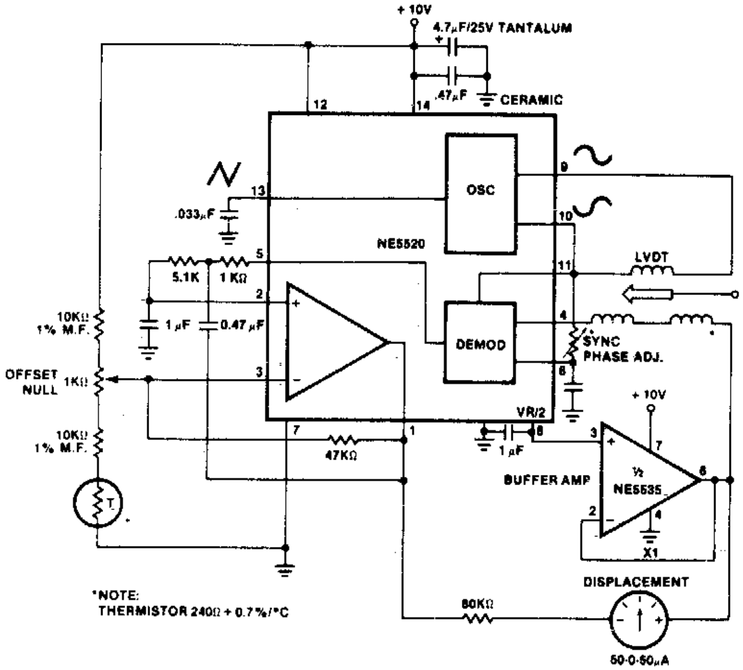

VR
VCC
+ 6V
.47μF
12
14
OSC
9
13
SINE WAVE CONVERTER
CT
0.033 μf
BIAS CKT
OSC
10
LVDT
NOTE
VR/2
GND
11
8
6
4
5.1 KΩ
1 KΩ
5
2
3
NE5520
VR/2
SYNCHRONOUS DEMODULATOR
A
C1
1 μF
C2
0.47 μF
R
4.7kΩ
LOW PASS FILTER
100K
Rf
47K
1
7
.47μF
– 6V
– VCC
10K
1K
10K
– VCC
+ VCC
OFFSET NULL
OUTPUT SIGNAL

Fig. 49-12

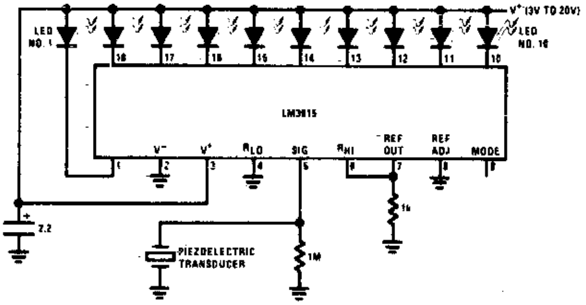
LINEAR VARIABLE DIFFERENTIAL TRANSFORMER (LVDT) MEASURING GAUGE



+10V
4.7μF/25V TANTALUM
.47μF
CERAMIC
12
14
13
.033μF
OSC
9
10
NE5520
5
5.1K
1 KΩ
2
1μF
0.47μF
LVDT
11
4
DEMOD
SYNC
PHASE ADJ.
8
10KΩ
1% M.F.
OFFSET
NULL
1KΩ
3
10KΩ
1% M.F.
7
47KΩ
1
VR/2
1μF
+10V
3
7
6
BUFFER AMP
½
NE5535
2
4
X1
T
*NOTE:
THERMISTOR 240Ω + 0.7%/°C
DISPLACEMENT
80KΩ
50-0-50μA

Fig. 49-13

VIBRATION METER



V⁺ (3V TO 20V)
LED NO. 1
LED NO. 10
18
17
16
15
14
13
12
11
10
LM3915
V⁻
V⁺
R_LO
SIG
R_HI
REF OUT
REF ADJ
MODE
1
2
3
4
5
6
7
8
9
2.2
PIEZOELECTRIC TRANSDUCER
1M
1k

LED	Threshold
1	60 mV
2	80 mV
3	110 mV
4	160 mV
5	220 mV
6	320 mV
7	440 mV
8	630 mV
9	890 mV
10	1.25 V

Fig. 49-14

SENSITIVE RF VOLTMETER

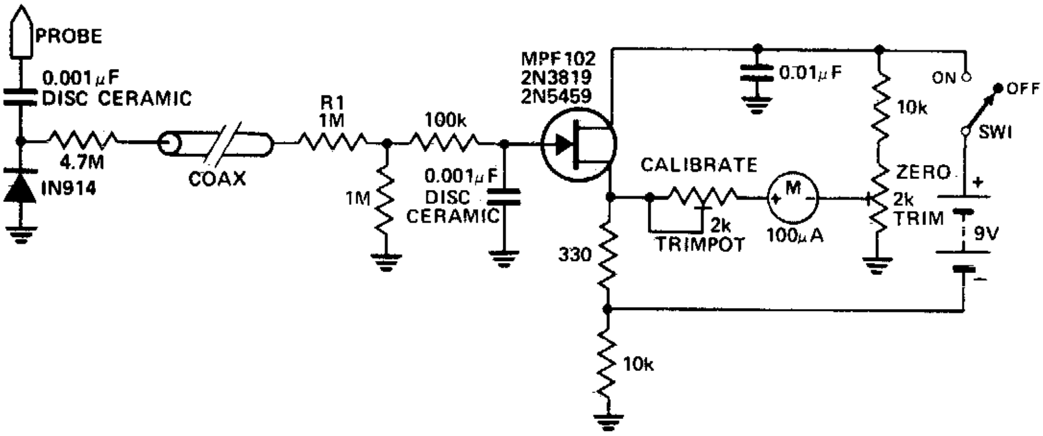

PROBE
0.001 μF
DISC CERAMIC
4.7M
IN914
COAX
R1
1M
1M
100k
0.001 μF
DISC
CERAMIC
MPF102
2N3819
2N5459
0.01 μF
10k
ON
OFF
SWI
CALIBRATE
M
2k
TRIMPOT
100 μA
ZERO
2k
TRIM
9V
330
10k

Fig. 49-15

Circuit Notes

This circuit measures RF voltages beyond 200 MHz and up to about 5 V. The diode should be mounted in a remote probe, close to the probe tip. Sensitivity is excellent and voltages less than 1 V peak can be easily measured. The unit can be calibrated by connecting the input to a known level of RF voltage, such as a calibrated signal generator, and setting the calibrate control.

MINIMUM COMPONENT TACHOMETER

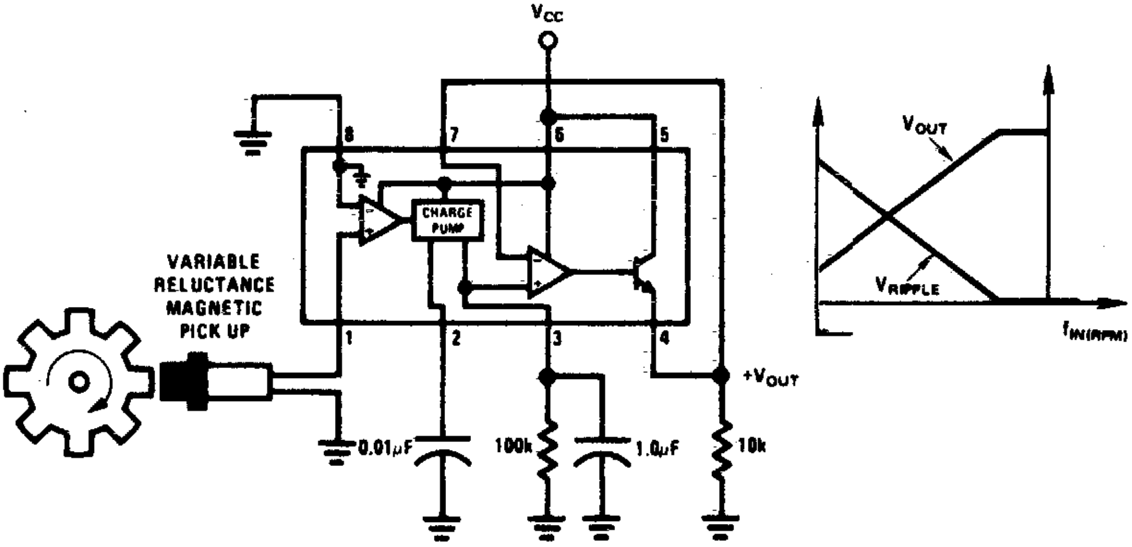
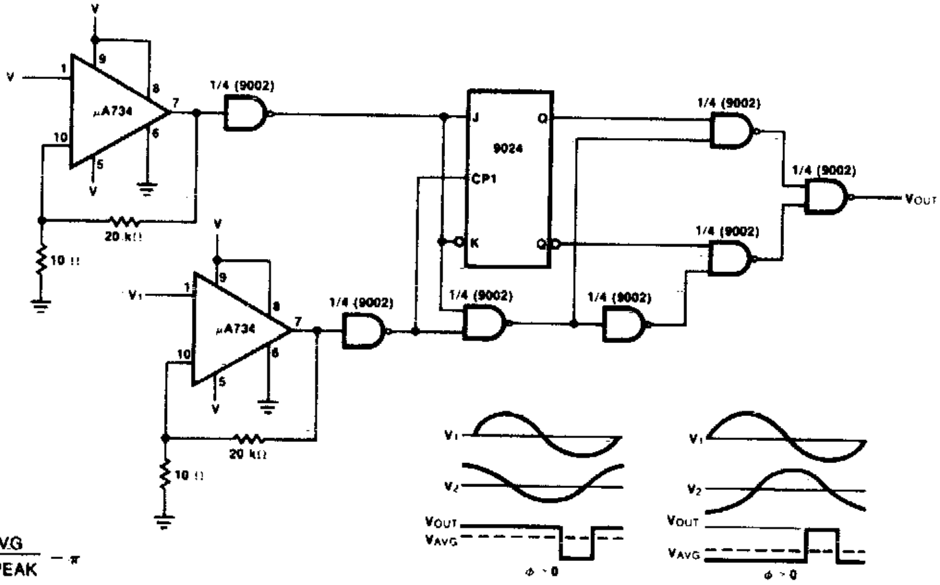

V_{CC}
8
7
6
5
CHARGE PUMP
1
2
3
4
VARIABLE RELUCTANCE MAGNETIC PICK UP
$+V_{OUT}$
0.01 μF
100k
1.0 μF
10k
V_{OUT}
V_{RIPPLE}
$f_{IN(RPM)}$

Fig. 49-16

PHASE METER


V
V
9
8
7
μA734
10
6
5
V
20 kΩ
10 Ω
1/4 (9002)
J
Q
9024
CP1
K
Q̄
1/4 (9002)
1/4 (9002)
1/4 (9002)
V_OUT
V
V₁
1
9
8
7
μA734
10
6
5
V
20 kΩ
10 Ω
1/4 (9002)
1/4 (9002)
1/4 (9002)
V₁
V₂
V_OUT
V_AVG
φ < 0
V₁
V₂
V_OUT
V_AVG
φ > 0

$$\phi = \frac{2\pi V_{AVG}}{V_{OUT,\ PEAK}} - \pi$$

Fig. 49-17

PRECISION CALIBRATION STANDARD

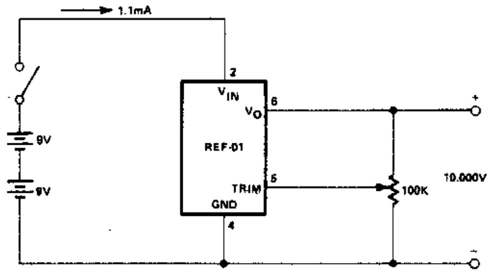

1.1mA
9V
9V
2
V_IN
V_O
6
REF-01
TRIM
5
GND
4
100K
+
10.000V
−

Fig. 49-18

Circuit Notes

An external power supply that gives a voltage higher than the highest expected rating of the zener diodes to be tested is required. Potentiometer RV1 is adjusted until the meter reading stabilizes. This reading is the zener diode's breakdown voltage.

ZENER DIODE CHECKER

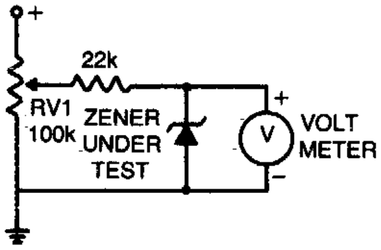

+
22k
RV1
100k
ZENER
UNDER
TEST
+
V
−
VOLT
METER

Fig. 49-19

50

Metal Detectors

The sources of the following circuits are contained in the Sources section beginning on page 730. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

Micropower Metal Detector

Lo-Parts Treasure Locator

MICROPOWER METAL DETECTOR

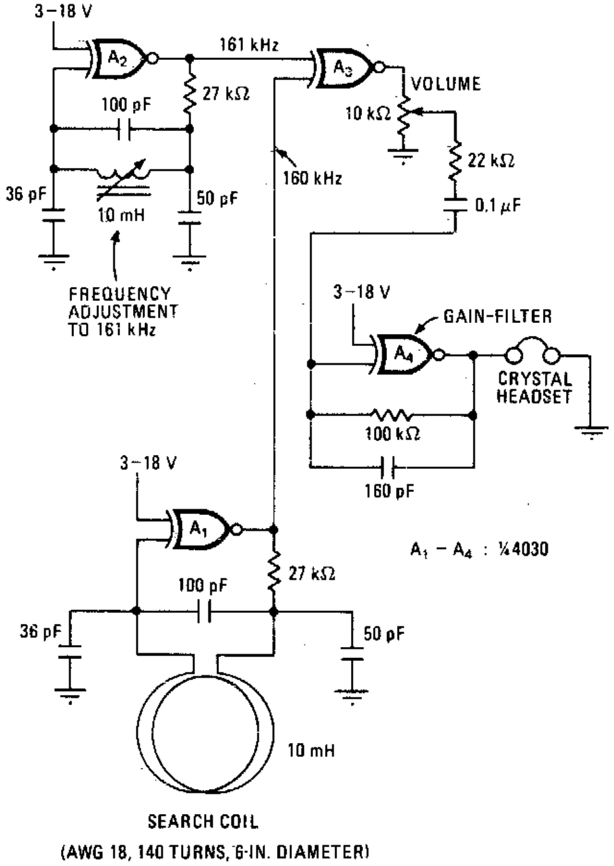

3–18 V
A2
161 kHz
A3
VOLUME
27 kΩ
100 pF
10 kΩ
22 kΩ
160 kHz
36 pF
10 mH
50 pF
0.1 μF
FREQUENCY ADJUSTMENT TO 161 kHz
3–18 V
GAIN-FILTER
A4
CRYSTAL HEADSET
100 kΩ
160 pF
3–18 V
A1
A1 – A4 : ¼4030
27 kΩ
100 pF
36 pF
50 pF
10 mH
SEARCH COIL
(AWG 18, 140 TURNS, 6-IN. DIAMETER)

Fig. 50-1

Circuit Notes

This battery-powered metal detector uses four exclusive-OR gates contained in the 4030 CMOS integrated circuit. The gates are wired as a twin-oscillators and a search coil serves as the inductance element in one of the oscillators. When the coil is brought near metal, the resultant change in its effective inductance changes the oscillator's frequency. Gates A1 and A2 form the two oscillators which are tuned to 160 and 161 kilohertz respectively. The pulses produced by each oscillator are mixed in A3, its output contains sum and difference frequencies at 1 and 321 kHz. The 321 kHz signal is filtered out by the 10 kHz low-pass filter at A4, leaving the 1 kHz signal to be amplified for the crystal headset connected at the output. The device's sensitivity is sufficient to detect coin-sized objects a foot away.

LO-PARTS TREASURE LOCATOR

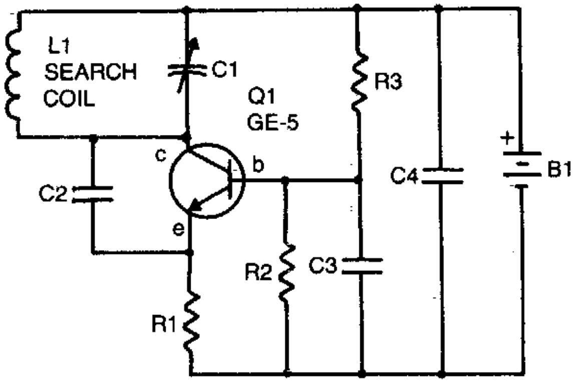


Fig. 50-2

PARTS LIST FOR
LO-PARTS TREASURE LOCATOR
B1—9-Vdc transistor battery
C1—365-pF trimmer or variable capacitor
C2—100-pF, 100-V silver mica capacitor
C3—0.05-μF, disc capacitor
C4—4.7- or 5-μF, 12-V electrolytic capacitor
L1—Search coil consisting of 18 turns of #22 enamel wire scramble wound on 4-in. diameter form
Q1—RCA SK3011 npn transistor or equiv.
R1—680-ohm, ½-watt resistor
R2—10,000-ohm, ½-watt resistor
R3—47,000-ohm, ½-watt resistor

Circuit Notes

Locator uses a transistor radio as the detector. With the radio tuned to a weak station, adjust C1 so the locator oscillator beats against the received signal. When the search head passes over metal, the inductance of L1 changes thereby changing the locator oscillator's frequency and changing the beat tone in the radio. The search coil consists of 18 turns of #22 enameled wire scramble wound on a 4-in. diameter form. After the coil is wound and checked for proper operation, saturate the coil with RTV adhesive for stable operation of the locator.

51

Metronomes

The sources of the following circuits are contained in the Sources section beginning on page 730. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

Accentuated Beat Metronome

Sight N' Sound Metronome

Micrometronome

ACCENTUATED BEAT METRONOME

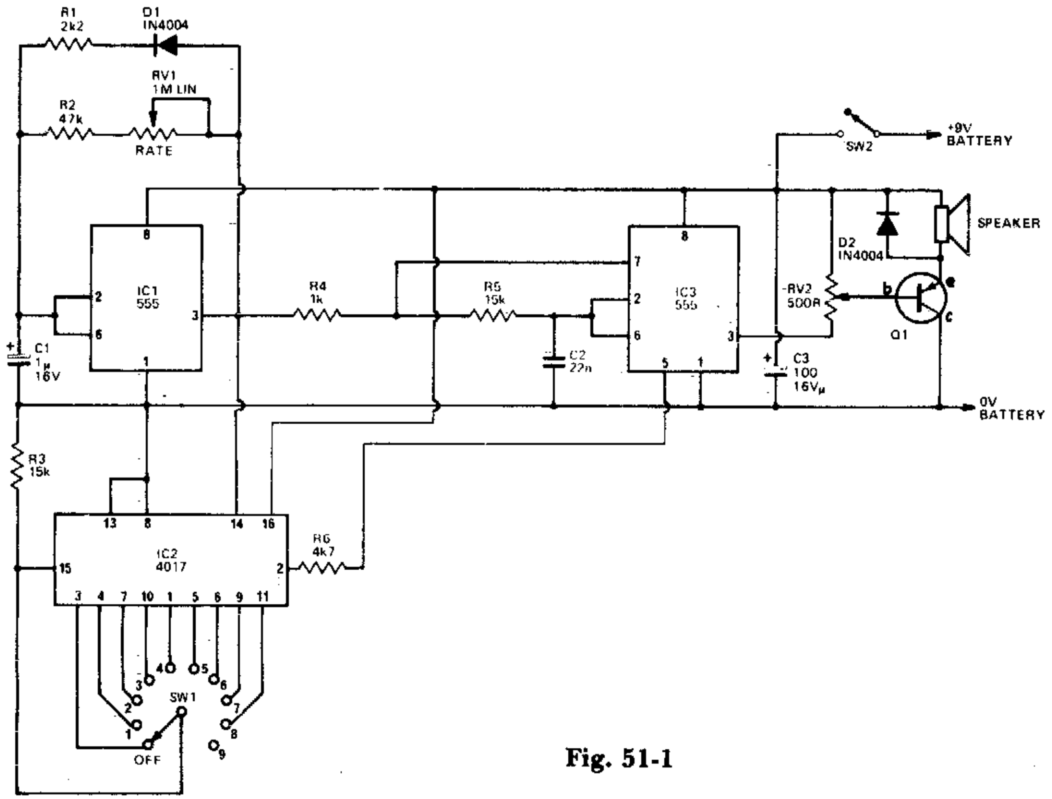

R1
2k2
D1
IN4004
RV1
1M LIN
R2
47k
RATE
+9V
BATTERY
SW2
SPEAKER
8
IC1
555
2
6
3
1
C1
1µ
16V
R4
1k
R5
15k
C2
22n
7
IC3
555
2
6
5
1
3
8
D2
IN4004
RV2
500R
b
e
c
Q1
C3
100
16Vµ
0V
BATTERY
R3
15k
13
8
14
16
IC2
4017
15
2
R6
4k7
3
4
7
10
1
5
6
9
11
SW1
1
2
3
4
5
6
7
8
9
OFF

Fig. 51-1

Circuit Notes

IC3 acts as an oscillator which operates if the output of IC1 is high. With the values used the two frequencies produced are about 800 Hz and 2500 Hz. The output is buffered by Q1 which drives the speaker. The first IC is used to generate the tone duration and the time interval between beats. The interval is adjustable by RV1 while the tone duration is set by R1. The output of IC1 also clocks IC2, a decade counter with 10 decoded outputs. Each of these outputs go high in sequence on each clock. The second output of IC2 is connected to the control input of IC3 and is used to change the frequency. Therefore the first tone will be high frequency, the second low and the third to tenth will be high again. This gives the 9-1 beat. If for example the 5th output is connected to the reset, the first tone will be high, the second low, and the third and fourth high, then when the 5th output goes to a high it resets it back to the first which is a high tone. We then have 3 high and one low tones or a 3-1.

SIGHT N' SOUND METRONOME

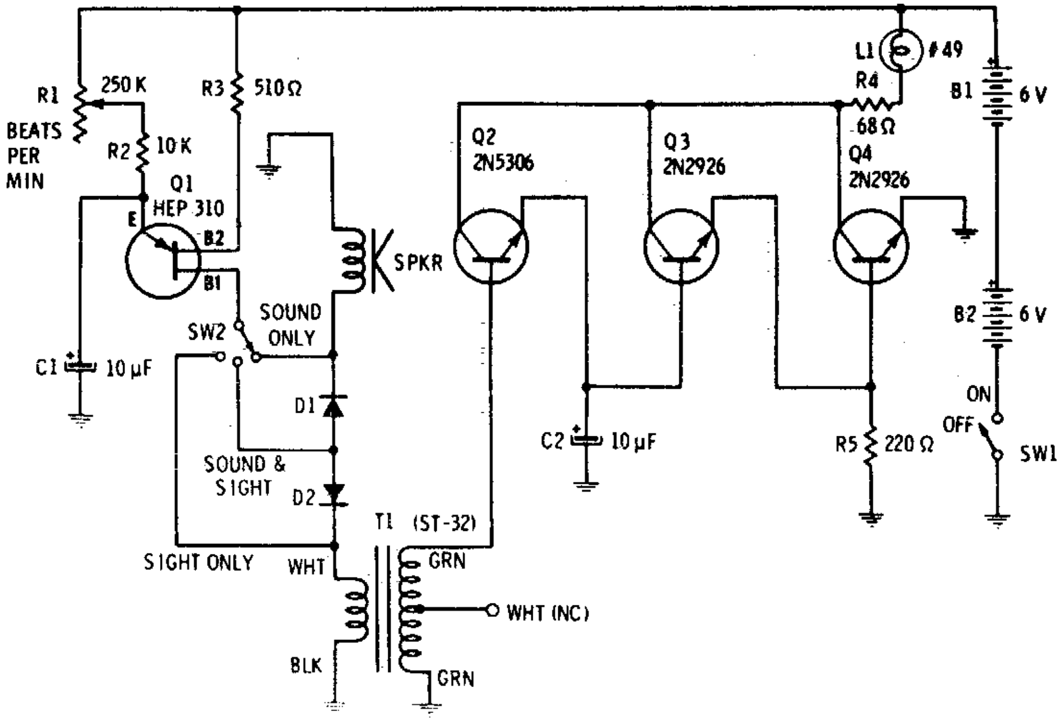

R1
250 K
BEATS
PER
MIN
R2
10 K
R3
510Ω
Q1
HEP 310
E
B2
B1
SPKR
SW2
SOUND
ONLY
C1
10 μF
D1
SOUND &
SIGHT
D2
SIGHT ONLY
WHT
BLK
T1 (ST-32)
GRN
WHT (NC)
GRN
Q2
2N5306
Q3
2N2926
Q4
2N2926
C2
10 μF
L1
#49
R4
68Ω
R5
220 Ω
B1
6 V
B2
6 V
ON
OFF
SW1

Fig. 51-2

Circuit Notes

Precise, adjustable control of beats per minute from a largo of 18 to a frenzied, high presto of 500. These beats are produced acoustically through a speaker. A light flashes at the same rate. When SW1 is closed, C1 begins to charge through R1 and R2. C1 will eventually reach a voltage at which the emitter of unijunction transistor is switched on, "dumping" the energy stored in C1 into an 8 ohm speaker. To produce a distinct "plop", brief pulses across T2 secondary drive Q2 into conduction. The extra gain of Q3 and Q4 are sufficient to briefly switch L1 on, then off, as the pulse wave passes. Capacitor C2 "stretches" the pulse slightly to overcome the thermal inertia of the lamp, so that a bright flash occurs.

MICROMETRONOME

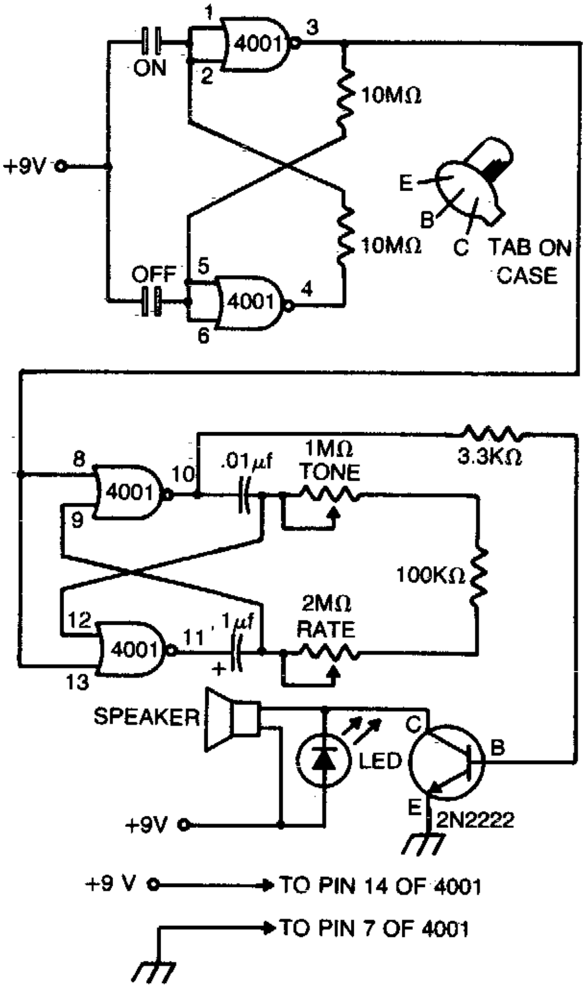

1
4001
3
ON
2
10MΩ
+9V
E
B
C
TAB ON CASE
10MΩ
OFF
5
4001
4
6
8
4001
10
.01μf
1MΩ
TONE
3.3KΩ
9
100KΩ
12
4001
11
1μf
+
2MΩ
RATE
13
SPEAKER
LED
C
B
E
2N2222
+9V
+9 V
TO PIN 14 OF 4001
TO PIN 7 OF 4001

Fig. 51-3

Circuit Notes

This compact metronome will run for years on a single nine-volt transistor battery. Has both tone and pulse rate controls, and uses touch plates to start and stop, can be built in a case no larger than a pack of cigarettes. The touch plates consist of two strips of metal about 1/16-inch apart mounted on, but insulated from, the case. Bridging the gap closes the switch.

52

Miscellaneous Circuits

The sources of the following circuits are contained in the Sources section beginning on page 730. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

Intercom
Musical Organ
Laser Diode Pulser
Capacitance Multiplier
Simulated Inductor
Active Inductor
Positive Input/Negative Output Charge Pump
Shift Register Driver
Tape Recorder
Negative-Edge Differentiator
Stylus Organ
Positive-Edge Differentiator
Four Channel Data Acquisition System
Triac Trigger
Precision Rectifiers
Voltage Control Resistor
Fast Inverter Circuit
Inverse Scaler
5.0 V Square Wave Calibrator
Low Drift Integrator and Low-Leakage Guarded Reset
Differentiator with High Common Mode Noise Rejection
Digital Transmission Isolator

INTERCOM

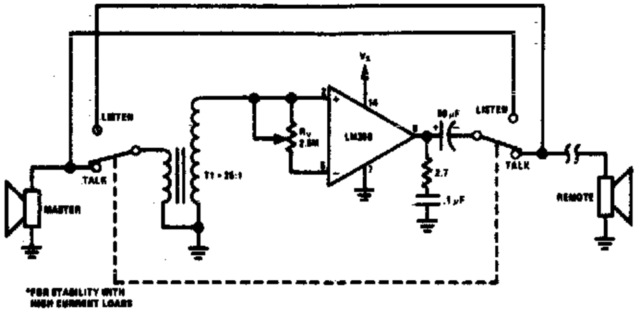

LISTEN
TALK
MASTER
T1 = 25:1
Rv
LM380
Vs
14
2
6
7
8
50 µF
LISTEN
2.7
.1 µF
TALK
REMOTE
*FOR STABILITY WITH HIGH CURRENT LOADS

Fig. 52-1

Circuit Notes

The circuit provides a minimum component intercom. With switch S1 in the talk position, the speaker of the master station acts as the microphone with the aid of step-up transformer T1. A turns ratio of 25 and a device gain of 50 allows a maximum loop gain of 1250. R_v provides a common mode volume control. Switching S1 to the listen position reverses the role of the master and remote speakers.

MUSICAL ORGAN

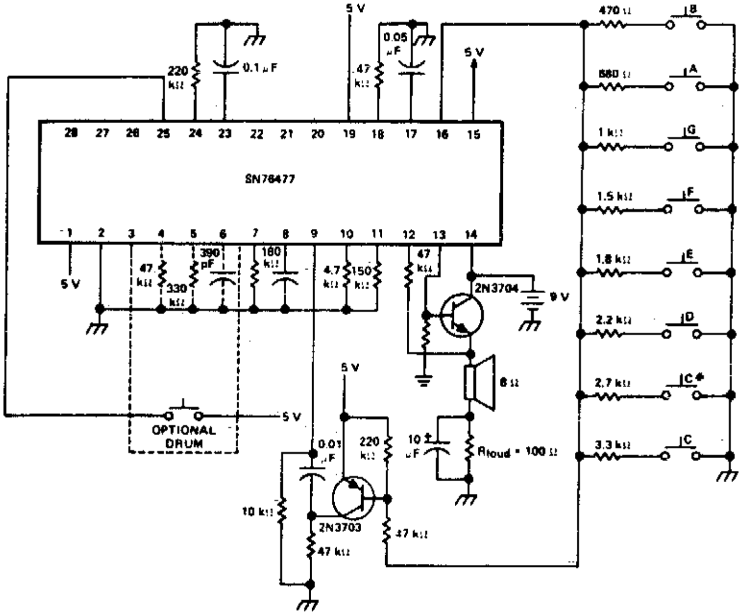

5 V
220 kΩ
0.1 µF
.47 kΩ
0.05 µF
5 V
28
27
26
25
24
23
22
21
20
19
18
17
16
15
SN76477
1
2
3
4
5
6
7
8
9
10
11
12
13
14
5 V
47 kΩ
330 kΩ
390 pF
180 kΩ
4.7 kΩ
150 kΩ
47 kΩ
2N3704
9 V
8 Ω
OPTIONAL DRUM
5 V
5 V
0.01 µF
220 kΩ
10 kΩ
2N3703
47 kΩ
47 kΩ
10 µF
Rloud = 100 Ω
470 Ω
B
680 Ω
A
1 kΩ
G
1.5 kΩ
F
1.8 kΩ
E
2.2 kΩ
D
2.7 kΩ
C#
3.3 kΩ
C

Fig. 52-2

LASER DIODE PULSER

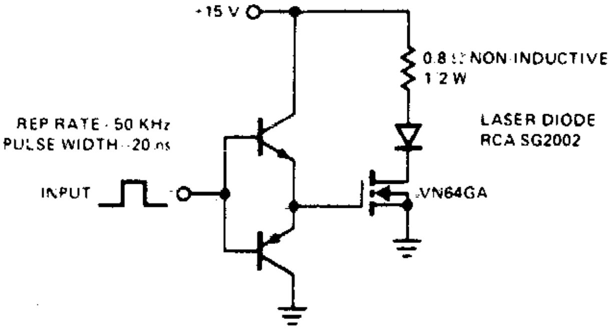
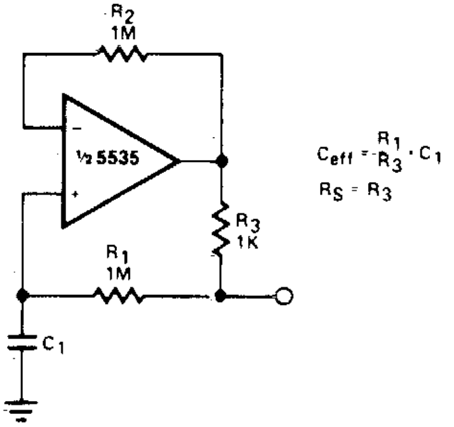

+15 V
0.8 Ω NON-INDUCTIVE
1/2 W
REP RATE - 50 KHz
PULSE WIDTH - 20 ns
LASER DIODE
RCA SG2002
INPUT
VN64GA

Fig. 52-3

Circuit Notes

This drive is capable of driving the laser-diode with 10 ampere, 20 ns pulses. For a 0.1% duty cycle, the repetition rate will be 50 kHz. A complementary emitter-follower is used as a driver. Switching speed is determined by the f_T of the bipolar transistors used and the impedance of the drive source.

CAPACITANCE MULTIPLIER


R2
1M
½ 5535
-
+
R3
1K
R1
1M
C1
$C_{eff} = \frac{R_1}{R_3} \cdot C_1$
$R_S = R_3$

All resistor values are in ohms

Fig. 52-4

Circuit Notes

This circuit can be used to simulate large capacitances using small value components. With the values shown and C = 10 μF, an effective capacitance of 10,000 μF was obtained. The Q available is limited by the effective series resistance. So R1 should be as large as practical.

SIMULATED INDUCTOR

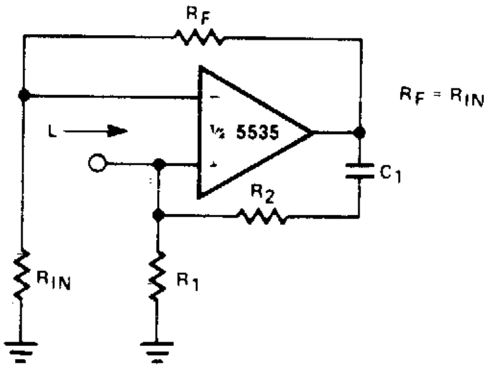
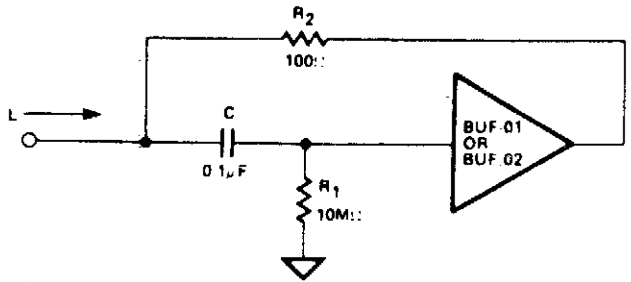


Fig. 52-5

Circuit Notes

With a constant current excitation, the voltage dropped across an inductance increases with frequency. Thus, an active device whose output increases with frequency can be characterized as an inductance. The circuit yields such a response with the effective inductance being equal to: L = R1R2C. The Q of this inductance depends upon R1 being equal to R2. At the same time, however, the positive and negative feedback paths of the amplifier are equal leading to the distinct possibility of instability at high frequencies. R1 should, therefore, always be slightly smaller than R2 to assure stable operation.

ACTIVE INDUCTOR



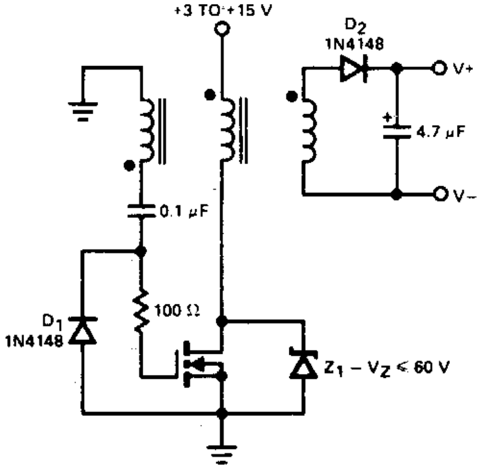
$L \geq R_1R_2C = 100$ HENRIES
$R_S = R_2 = 100\Omega$
$R_P = R_1 = 10$ MEGΩ
ASSUMING C_{STRAY} (ACROSS R_1) OF 5 pF THE UPPER
FREQUENCY LIMIT IS APPROXIMATELY 7kHz.
$X_L = 100\Omega$ AT $f = 0.159$Hz

Fig. 52-6

Circuit Notes

An active inductor is realized with an eight-lead IC, two carbon resistors, and a small capacitor. A commercial inductor of 50 henries may occupy up to five cubic inches.

POSITIVE INPUT/NEGATIVE OUTPUT CHARGE PUMP


+3 TO +15 V
D2
1N4148
V+
4.7 μF
V−
0.1 μF
100 Ω
D1
1N4148
Z1 − Vz ≤ 60 V

TRANSFORMER:
INDIANA GENERAL CORE F626-12-Q2
26 TURNS NO. 28 WIRE TRIFILAR WOUND

Circuit Notes

A simple means of generating a low-power voltage supply of opposite polarity from the main supply. Self oscillating driver produces pulses at a repetition frequency of 100 kHz. When the VMOS device is off, capacitor C is charged to the positive supply. When the VMOS transistor switches on, C delivers a negative voltage through the series diode to the output. The zener serves as a dissipative regulator.

Fig. 52-7

SHIFT REGISTER DRIVER

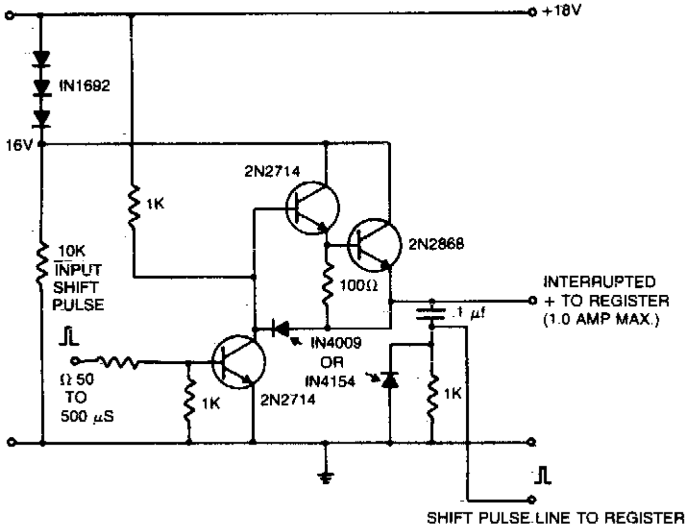
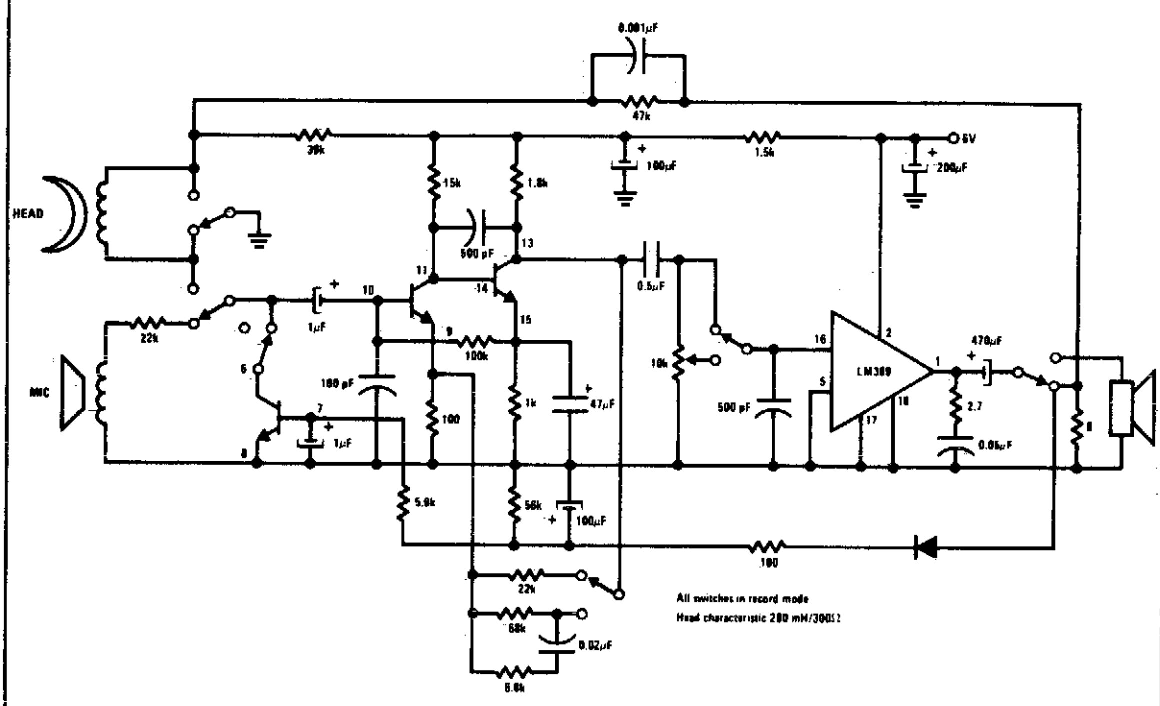

+18V
IN1692
16V
2N2714
1K
2N2868
10K
INPUT
SHIFT
PULSE
100Ω
INTERRUPTED
+ TO REGISTER
(1.0 AMP MAX.)
.1 μf
IN4009
OR
IN4154
Ω 50
TO
500 μS
1K
2N2714
1K
SHIFT PULSE LINE TO REGISTER

Fig. 52-8

Circuit Notes

A 16 V power supply can be synthesized as shown using IN1692 rectifiers. A shift pulse input saturates the 2N2714 depriving the Darlington combination (2N2714 and 2N2868) of base drive. The negative pulse so generated on the 15 V line is differentiated to produce a positive trigger pulse at its trailing edge.

TAPE RECORDER


0.001µF
47k
39k
15k
1.8k
100µF
1.5k
6V
200µF
HEAD
500pF
13
11
14
10
1µF
0.5µF
15
9
100k
22k
MIC
6
10k
16
2
LM389
470µF
1
5
18
17
2.7
0.05µF
8
7
1µF
100pF
100
1k
47µF
500pF
5.6k
56k
100µF
100
22k
68k
0.02µF
6.8k

All switches in record mode

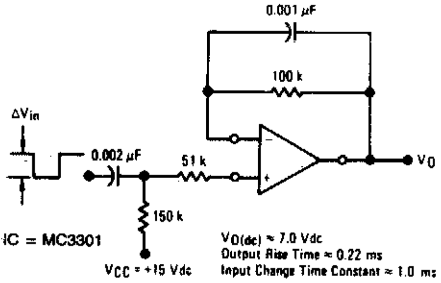
Head characteristic 200 mH/300Ω

Fig. 52-9

Circuit Notes

Complete record/playback cassette tape machine amplifier. Two of the transistors act as signal amplifiers, with the third used for automatic level control during the record mode.

NEGATIVE-EDGE DIFFERENTIATOR


0.001 µF
100 k
ΔVin
0.002 µF
51 k
150 k
VO

IC = MC3301

V_{CC} = +15 Vdc

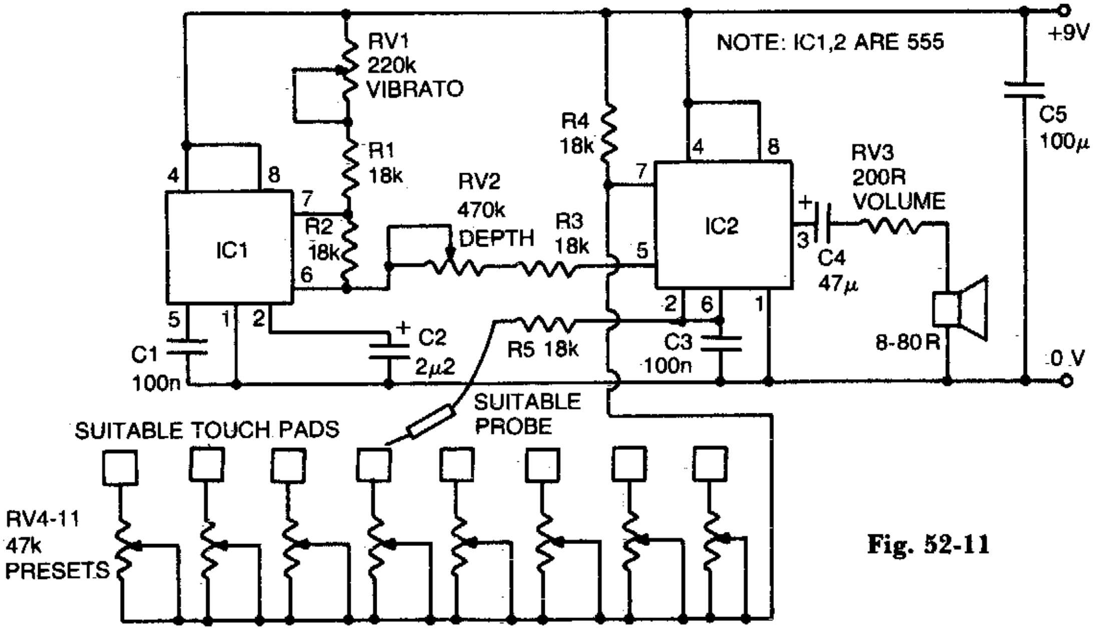
$V_{O(dc)}$ ≈ 7.0 Vdc

Output Rise Time ≈ 0.22 ms

Input Change Time Constant ≈ 1.0 ms

Fig. 52-10

STYLUS ORGAN



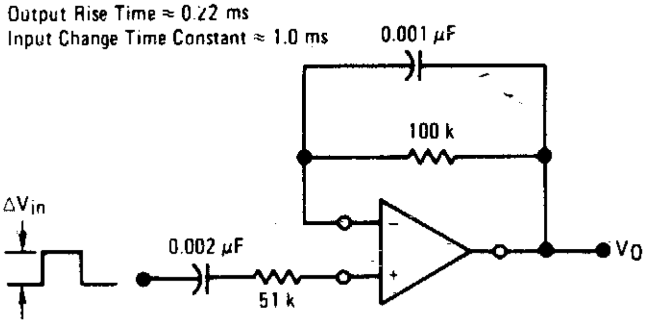
NOTE: IC1,2 ARE 555
+9V
0 V
RV1
220k
VIBRATO
R1
18k
R2
18k
RV2
470k
DEPTH
R3
18k
R4
18k
IC1
IC2
4
8
7
6
5
1
2
3
C1
100n
C2
2μ2
C3
100n
C4
47μ
C5
100μ
RV3
200R
VOLUME
8-80R
R5 18k
SUITABLE
PROBE
SUITABLE TOUCH PADS
RV4-11
47k
PRESETS

Fig. 52-11

Circuit Notes

IC2 is an audio frequency oscillator. Its frequency is primarily controlled by the resistance between pins 2 and 7. RV4-11 control the oscillator frequency and by touching a stylus (connected via limiting resistor R5 to pin 2) to each preset, different notes can be played. IC1 is a low frequency oscillator (approximately 3-10Hz), the frequency of which is variable by RV1. The output of this oscillator is connected through depth control RV2 and limiting resistor R3 to the voltage control input of the audio frequency oscillator. Thus a vibrato effect occurs.

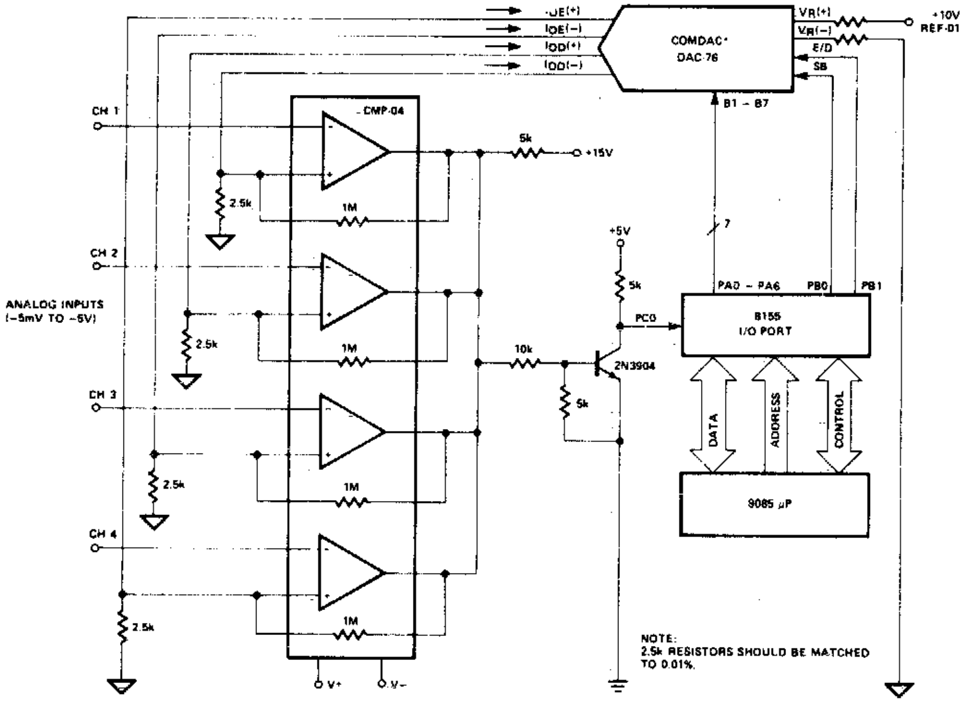
POSITIVE-EDGE DIFFERENTIATOR



Output Rise Time ≈ 0.22 ms
Input Change Time Constant ≈ 1.0 ms
0.001 μF
100 k
ΔVin
0.002 μF
51 k
−
+
V0

Fig. 52-12

FOUR CHANNEL DATA ACQUISITION SYSTEM

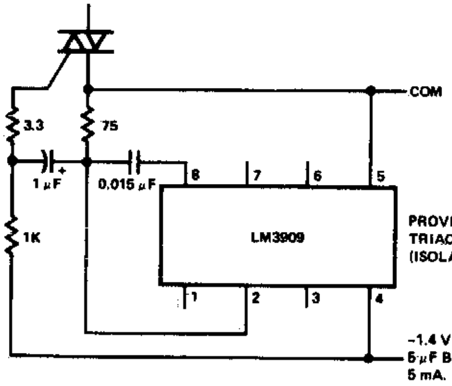


IOE(+)
IOE(−)
IOD(+)
IOD(−)
COMDAC®
DAC-76
VR(+)
VR(−)
E/D
SB
+10V
REF-01
B1 – B7
CMP-04
CH 1
CH 2
CH 3
CH 4
ANALOG INPUTS
(−5mV TO −5V)
2.5k
1M
5k
+15V
10k
+5V
5k
PCO
2N3904
5k
7
PA0 – PA6
PB0
PB1
8155
I/O PORT
DATA
ADDRESS
CONTROL
8085 μP
V+
V−

NOTE:
2.5k RESISTORS SHOULD BE MATCHED TO 0.01%.

Fig. 52-13

TRIAC TRIGGER



COM
3.3
75
1 μF
0.015 μF
1K
LM3909
8
7
6
5
1
2
3
4

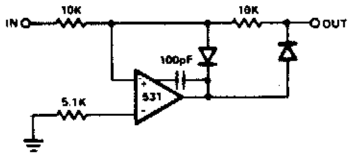
PROVIDES 40 mA; 10 μs PULSES AT ABOUT 8 kHz.
TRIAC GATE MAY BE PULSE TRANSFORMER
(ISOLATED IF DESIRED).

−1.4 V FROM BATTERY OR SOLAR CELL WITH
5 μF BYPASS CAPACITOR, DRAIN NOMINALLY
5 mA.

Fig. 52-14

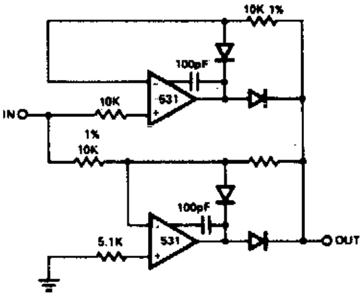
PRECISION RECTIFIERS

(a) HALF WAVE



IN
10K
10K
OUT
100pF
531
5.1K

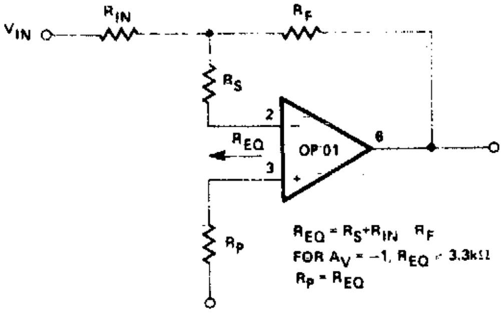
(b) FULL WAVE



10K 1%
100pF
531
IN
10K
1%
10K
100pF
5.1K
531
OUT

Fig. 52-15

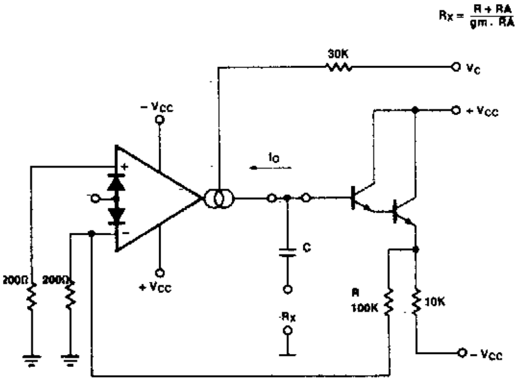
FAST INVERTER CIRCUIT



V_{IN}
R_{IN}
R_F
R_S
2
R_{EQ}
OP 01
6
3
R_P
$R_{EQ} = R_S + R_{IN} \parallel R_F$
FOR $A_V = -1$, $R_{EQ} \approx 3.3k\Omega$
$R_P = R_{EQ}$

Fig. 52-17

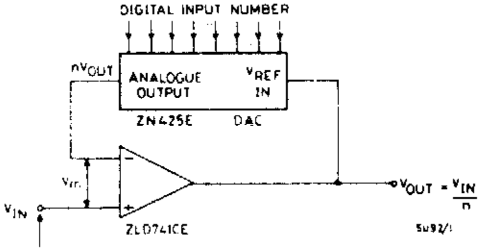
VOLTAGE CONTROL RESISTOR



$R_X = \frac{R + RA}{gm \cdot RA}$
30K
V_C
$-V_{CC}$
$+V_{CC}$
I_O
C
200Ω
200Ω
$+V_{CC}$
R
100K
10K
R_X
$-V_{CC}$

Fig. 52-16

INVERSE SCALER



DIGITAL INPUT NUMBER
nV_{OUT}
ANALOGUE OUTPUT
V_{REF} IN
ZN425E
DAC
V_{IN}
V_{IN}
ZLD741CE
$V_{OUT} = \frac{V_{IN}}{n}$
5492/1

Fig. 52-18

Circuit Notes

If a DAC is operated in the feedback loop of an operational amplifier, then the amplifier gain is inversely proportional to the input digital number or code to the DAC. The version giving scaling inversely proportional to positive voltage is shown.

5.0 V SQUARE WAVE CALIBRATOR

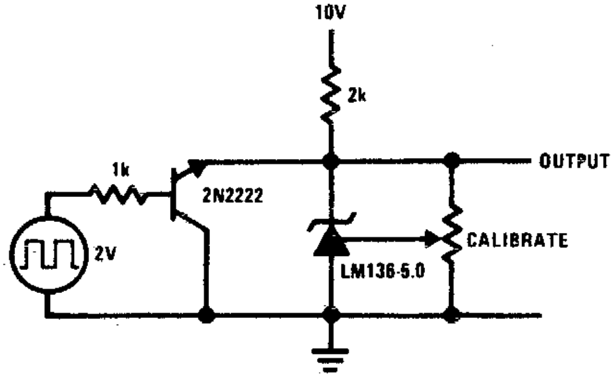
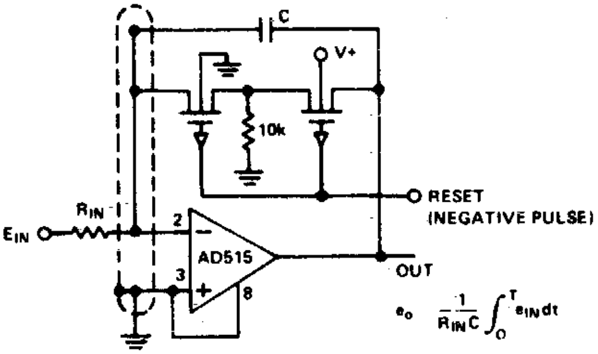

10V
2k
1k
2N2222
2V
OUTPUT
CALIBRATE
LM136-5.0

Fig. 52-19

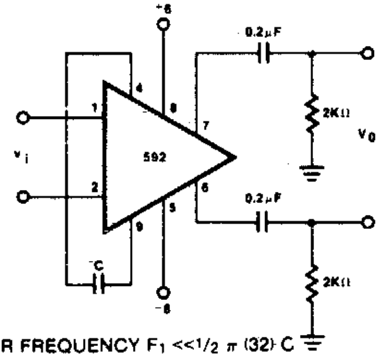
LOW DRIFT INTEGRATOR AND LOW-LEAKAGE GUARDED RESET


C
V+
10k
RESET
(NEGATIVE PULSE)
R_{IN}
E_{IN}
2
3
8
AD515
OUT

$e_o = -\frac{1}{R_{IN}C}\int_0^T e_{IN}\,dt$

Fig. 52-20

DIFFERENTIATOR WITH HIGH COMMON MODE NOISE REJECTION


+6
0.2μF
4
1
8
7
v_i
592
2KΩ
V_0
2
6
5
9
0.2μF
C
2KΩ
−6

FOR FREQUENCY $F_1 << 1/2\pi(32)C$
$V_0 \cong 1.4 \times 10^4 C \frac{dVi}{dT}$

Fig. 52-21

DIGITAL TRANSMISSION ISOLATOR

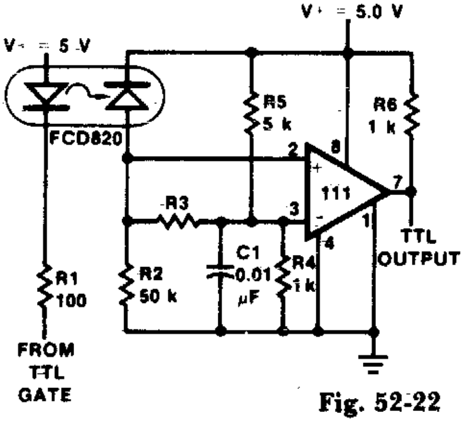

V+ = 5.0 V
V+ = 5 V
FCD820
R5
5 k
R6
1 k
2
8
7
111
R3
3
1
4
C1
0.01
μF
R4
1 k
R2
50 k
R1
100
TTL
OUTPUT
FROM
TTL
GATE

Fig. 52-22

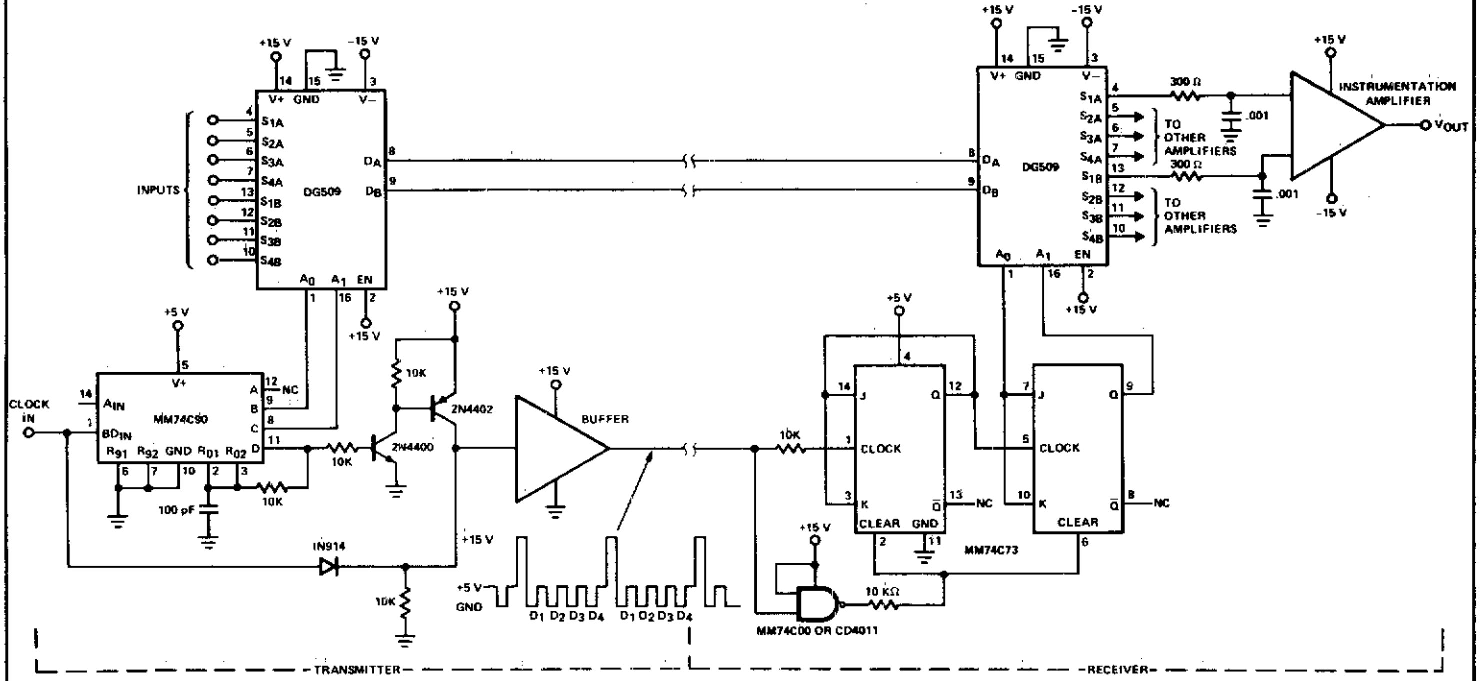
53

Mixers and Multiplexers

The sources of the following circuits are contained in the Sources section beginning on page 730. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

Differential Mux/Demux System
Eight Channel Mux/Demux System
Doubly Balanced Mixer
Common-Source Mixer
100 MHz Mixer
Multiplexer/Mixer
Wide Band Differential Multiplexer

DIFFERENTIAL MUX/DEMUX SYSTEM



INPUTS
DG509
MM74C90
CLOCK IN
+15 V
-15 V
+5 V
10K
100 pF
IN914
2N4402
2N4400
BUFFER
+5 V
GND
D1 D2 D3 D4
TRANSMITTER
RECEIVER
MM74C73
MM74C00 OR CD4011
10 KΩ
CLOCK
CLEAR
NC
TO OTHER AMPLIFIERS
300 Ω
.001
INSTRUMENTATION AMPLIFIER
V_{OUT}

Fig. 53-1

EIGHT CHANNEL MUX/DEMUX SYSTEM

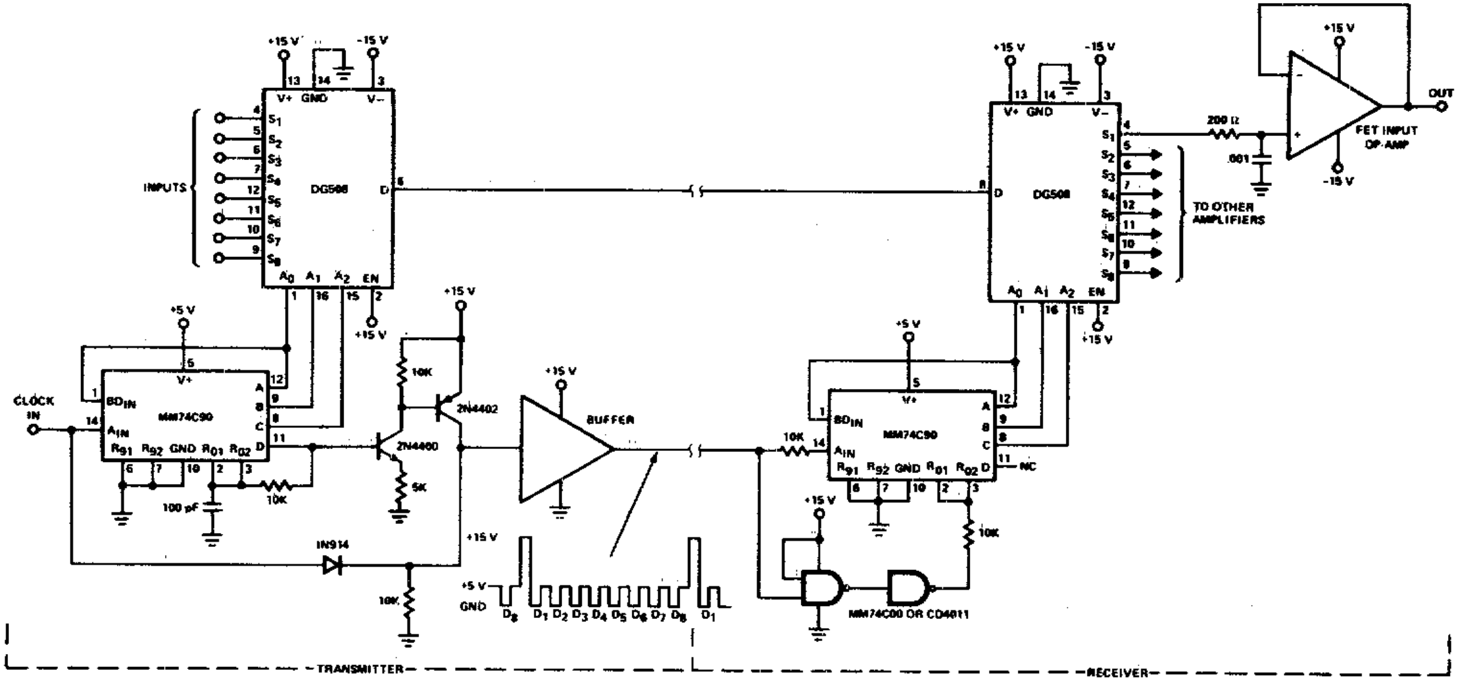
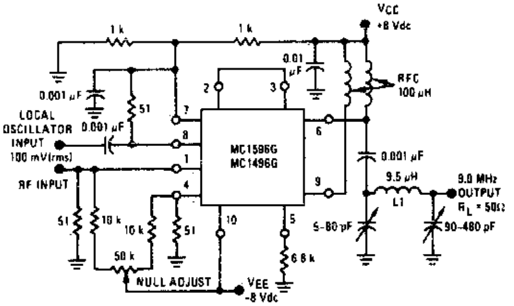

+15 V
-15 V
V+
GND
V–
INPUTS
S1
S2
S3
S4
S5
S6
S7
S8
DG508
D
A0
A1
A2
EN
+5 V
CLOCK IN
BDIN
AIN
MM74C90
R91
R92
GND
R01
R02
100 pF
10K
2N4402
2N4400
5K
IN914
BUFFER
GND
D8
D1 D2 D3 D4 D5 D6 D7 D8
D1
MM74C00 OR CD4011
NC
TO OTHER AMPLIFIERS
200 Ω
.001
FET INPUT OP-AMP
OUT
TRANSMITTER
RECEIVER

Fig. 53-2

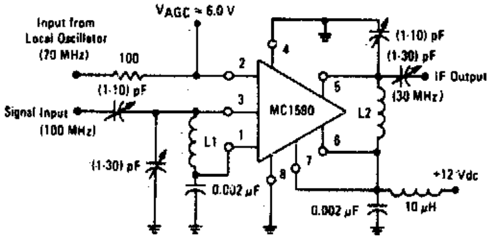
DOUBLY BALANCED MIXER (BROADBAND INPUTS, 9.0 MHz TUNED OUTPUT)


VCC
+8 Vdc
1 k
1 k
0.01 μF
RFC
100 μH
0.001 μF
51
LOCAL
OSCILLATOR
INPUT
100 mV(rms)
0.001 μF
RF INPUT
MC1596G
MC1496G
2
3
7
8
1
4
6
9
10
5
0.001 μF
9.5 μH
L1
9.0 MHz
OUTPUT
R_L = 50Ω
5-80 pF
90-480 pF
51
10 k
10 k
51
50 k
NULL ADJUST
VEE
-8 Vdc
6.8 k

L1 = 44 TURNS AWG NO. 28 ENAMELED WIRE, WOUND ON MICROMETALS TYPE 44-6 TOROID CORE.

Fig. 53-3

100 MHz MIXER

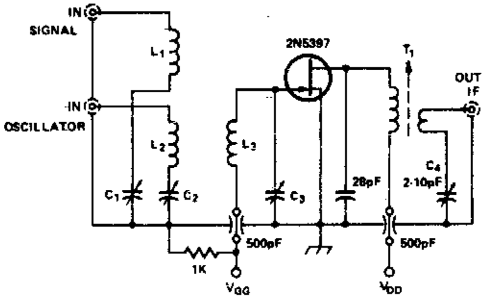

V_{AGC} ≈ 6.0 V
Input from
Local Oscillator
(70 MHz)
100
(1-10) pF
Signal Input
(100 MHz)
(1-30) pF
L1
0.002 μF
MC1590
2
3
1
4
5
6
7
8
(1-10) pF
(1-30) pF
IF Output
(30 MHz)
L2
+12 Vdc
0.002 μF
10 μH

L1 = 5 Turns, #16 AWG Wire, 1/4" ID, 5/8" Long

L2 = 16 Turns, #20 AWG Wire on a Toroid Core, (T44-6 Micro Metal or Equiv)

Fig. 53-5

COMMON-SOURCE MIXER


IN
SIGNAL
IN
OSCILLATOR
L_1
L_2
L_3
C_1
C_2
C_3
2N5397
T_1
OUT
IF
28pF
C_4
2-10pF
500pF
500pF
1K
V_{GG}
V_{DD}

L_1 – 1.4" long; #22 enamel, close coupled to L_3
L_2 – 1.6" long; #22 enamel, close coupled to L_3
L_3 – 1.75" long; #18 copper
T_1 – Primary, 13T; #22 enamel, close wound on 1/4" form ≅ 1μH
Secondary, 3T; #22 enamel, close wound over primary
C_1, C_2, C_3, C_4, – 0.8 to 12pF; Johanson type 2950

Fig. 53-4

MULTIPLEXER/MIXER

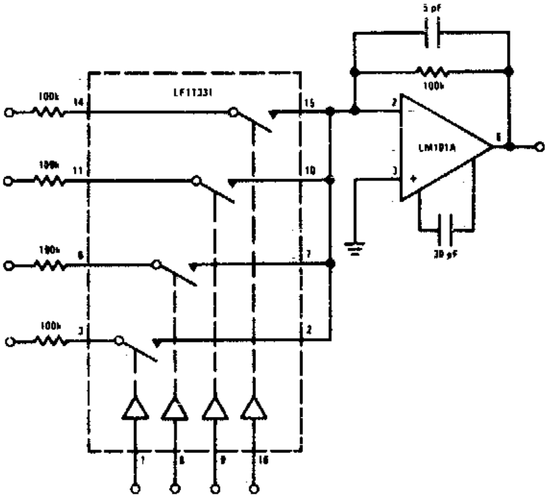

5 pF
100k
LF11331
100k
14
15
100k
11
10
100k
6
7
100k
3
2
LM101A
2
3
6
30 pF
1
8
9
16

Fig. 53-6

WIDE BAND DIFFERENTIAL MULTIPLEXER

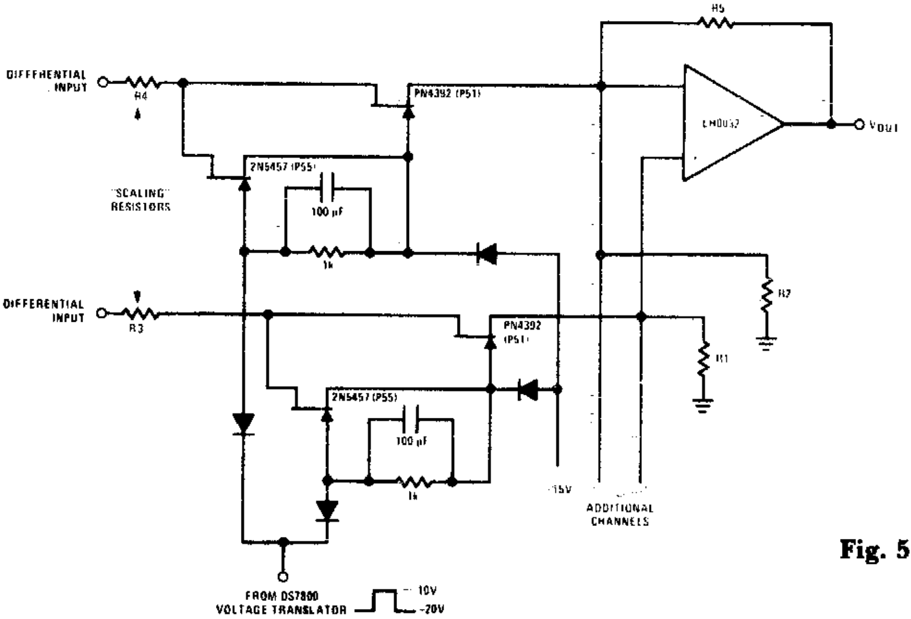

R5
DIFFERENTIAL INPUT
R4
PN4392 (P51)
LH0032
VOUT
2N5457 (P55)
"SCALING" RESISTORS
100 pF
1k
R2
DIFFERENTIAL INPUT
R3
PN4392 (P51)
R1
2N5457 (P55)
100 pF
1k
-15V
ADDITIONAL CHANNELS
FROM DS7800 VOLTAGE TRANSLATOR
10V
-20V

Fig. 53-7

Circuit Notes

This design allows high frequency signal handling and high toggle rates simultaneously. Toggle rates up to 1 MHz and MHz signals are possible with this circuit.

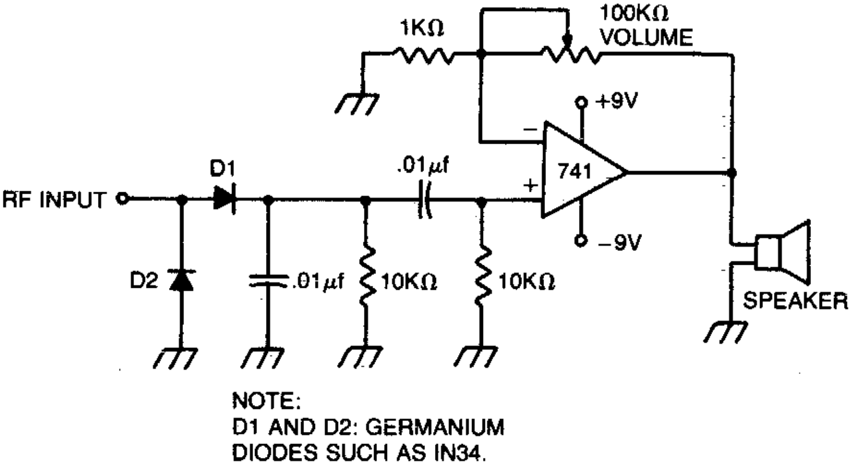
54

Modulation Monitors

The sources of the following circuits are contained in the Sources section beginning on page 730. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

Modulation Monitor
Visual Modulation Indicator
CB Modulation Monitor

MODULATION MONITOR



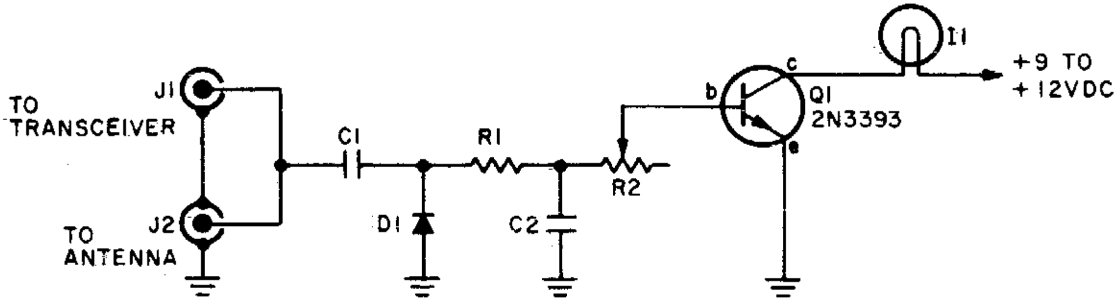
1KΩ
100KΩ
VOLUME
+9V
741
−9V
+
−
D1
.01μf
RF INPUT
D2
.01μf
10KΩ
10KΩ
SPEAKER
NOTE:
D1 AND D2: GERMANIUM
DIODES SUCH AS IN34.

Fig. 54-1

Circuit Notes

Broad-tuned receiver demodulates the RF signal picked up by a loosely coupled wire placed near the transmitting antenna.

VISUAL MODULATION INDICATOR



I1
+9 TO
+12VDC
c
b
e
Q1
2N3393
TO
TRANSCEIVER
J1
C1
R1
R2
TO
ANTENNA
J2
D1
C2

Fig. 54-2

Circuit Notes

Indicator lamp brightness varies in step with modulated RF signal. Adjust R2 with transmitter on (modulated) until the lamp flashes in step with modulation. C1 = 5 pf, C2 = 100 pF, D1 = 1N60 or 1N34 (Germanium), R3 = 10 K pot, I1 = 6-8 V, 30-60 mA incandescent bulb, Q1 = 2N3393 (for increased sensitivity use 2N3392 or other high-gain transistor).

CB MODULATION MONITOR

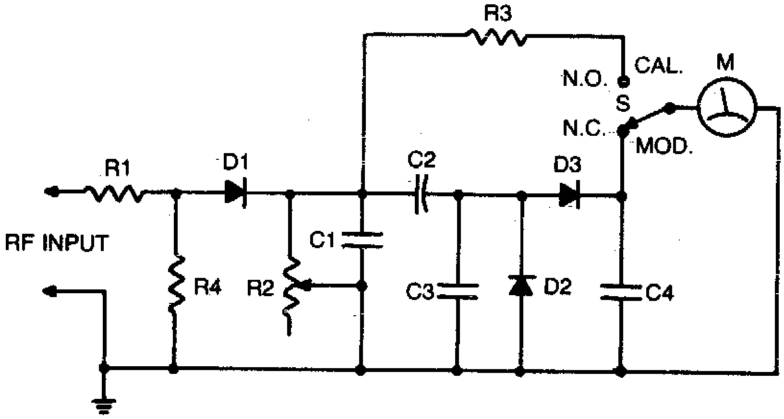


Fig. 54-3

PARTS LIST

C1—500-pF, 100-Vdc capacitor
C2—10-μF, 10-Vdc electrolytic capacitor
C3—200-pF, 100-Vdc capacitor
C4—300-pF, 100-Vdc capacitor
D1, D2, D3—1N60
M1—0-1 mA DC high-speed meter
R1, R4—1000-ohm, ½-watt resistor
R2—1000-ohm pot
R3—910-ohm, ½-watt resistor, 5%
S1—Spdt spring-return switch

Circuit Notes

Connect this circuit to a transceiver with a coaxial T connector in the transmission line. Key the transmitter (unmodulated), set S1 to CAL, and adjust R2 for a full scale reading. Return S1 to MOD position. The meter will read % modulation with 10% accuracy.

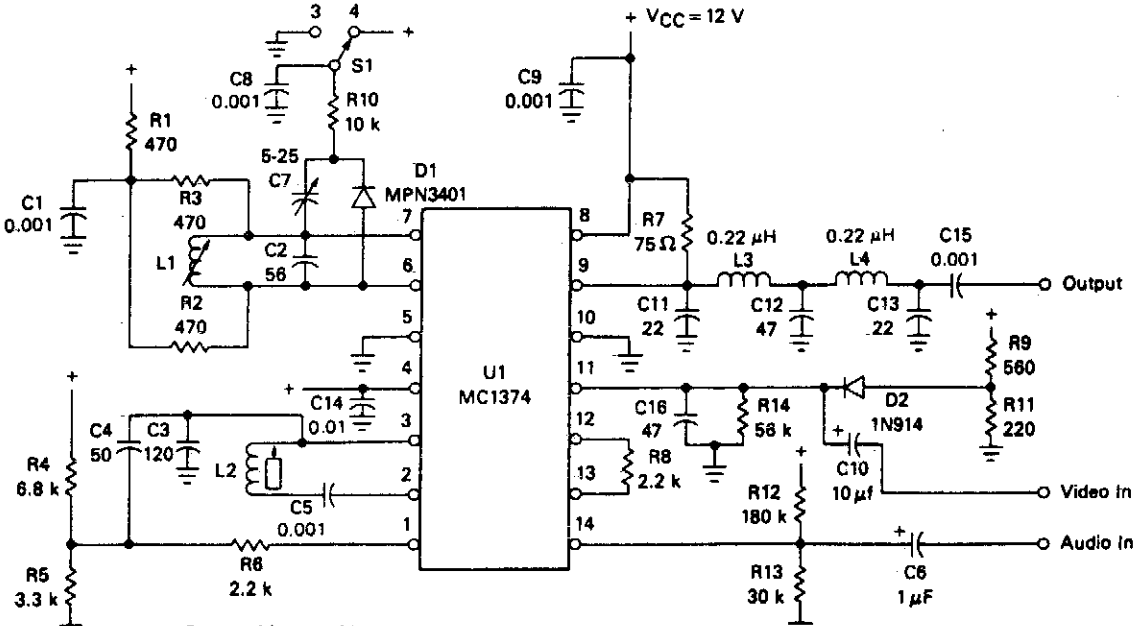
55

Modulators

The sources of the following circuits are contained in the Sources section beginning on page 730. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

TV Modulator
TV Modulator
Pulse-Position Modulator
Pulse-Width Modulator
Pulse-Width Modulator
RF Modulator
Linear Pulse-Width Modulator
Balanced Modulator
Video Modulator
Modulator
Pulse-Width Modulator
AM Modulator
TV Modulator Using a Motorola MC1374
Pulse-Width Modulator
Pulse-Width Modulator
VHF Modulator

TV MODULATOR


3
4
S1
+
C8
0.001
R10
10 k
+ VCC = 12 V
C9
0.001
+
R1
470
5-25
C7
D1
MPN3401
C1
0.001
R3
470
L1
C2
56
R2
470
7
6
5
4
3
2
1
8
9
10
11
12
13
14
U1
MC1374
R7
75 Ω
0.22 µH
L3
0.22 µH
L4
C15
0.001
Output
C11
22
C12
47
C13
22
+
R9
560
R11
220
D2
1N914
C16
47
R14
56 k
R8
2.2 k
+
C10
10 µf
Video In
R12
180 k
R13
30 k
+
C6
1 µF
Audio In
+
C14
0.01
+
C4
50
C3
120
L2
C5
0.001
R4
6.8 k
R5
3.3 k
R6
2.2 k

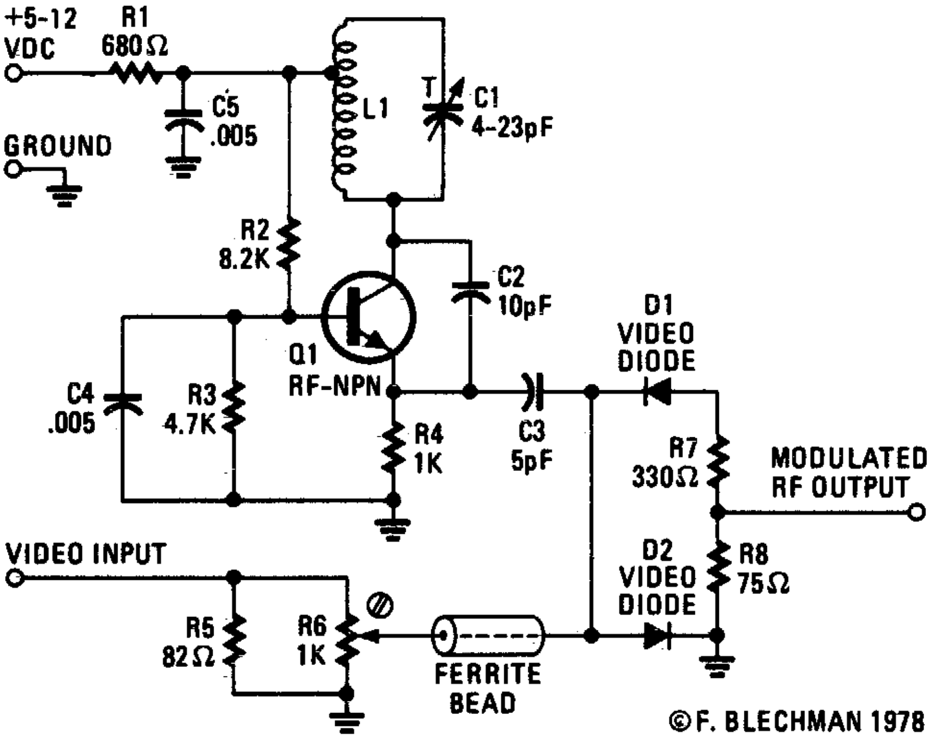
L1 - 4 Turns, #22, 1/4" Dia.
L2 - 40 Turns, #36, 3/16" Dia.

Fig. 55-1

Circuit Notes

The FM oscillator/modulator is a voltage-controlled oscillator, which exhibits a nearly linear output frequency versus input voltage characteristic for a wide deviation. It provides a good FM source with a few inexpensive external parts. It has a frequency range of 1.4 to 14 MHz and can typically produce a ±25 kHz modulated 4.5 MHz signal with about 0.6% total harmonic distortion.

TV MODULATOR



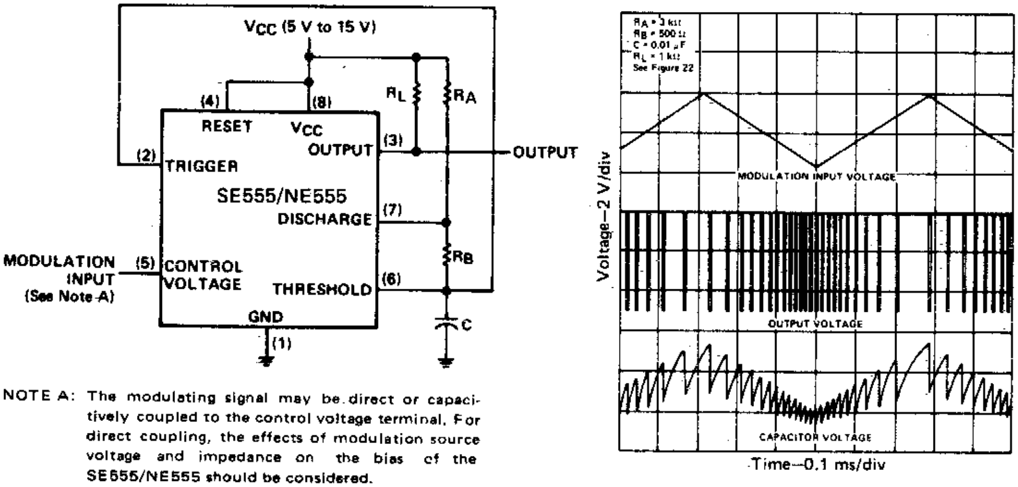
+5-12 VDC
R1 680Ω
C5 .005
GROUND
L1
T
C1 4-23pF
R2 8.2K
C2 10pF
Q1 RF-NPN
C4 .005
R3 4.7K
R4 1K
C3 5pF
D1 VIDEO DIODE
R7 330Ω
MODULATED RF OUTPUT
VIDEO INPUT
R5 82Ω
R6 1K
FERRITE BEAD
D2 VIDEO DIODE
R8 75Ω
©F. BLECHMAN 1978

Fig. 55-2

Circuit Notes

The VHF frequency is generated by a tuned Hartley oscillator circuit. Resistors R2, R3, and R4 bias the transistor, with tapped inductor L1 and trimmer capacitor C1 forming the tank circuit. Adjusting C1 determines the frequency. Capacitor C2 provides positive feedback from the tank circuit to the emitter at Q1. Capacitor C4 provides an RF ground for the base of Q1. Bypass capacitor C5 and resistor R1 filter out the radio frequencies generated in the tank circuit to prevent radiation from the power-supply lines. The video signal enters the parallel combination of resistors R5 and R6; this combination closely matches the 75 ohm impedance of most video cables. Resistor R6 is a small screwdriver-adjusted potentiometer that is used to control the video input level to mixer diodes D1 and D2.

PULSE-POSITION MODULATOR



VCC (5 V to 15 V)
(4)
(8)
RL
RA
RESET
VCC
OUTPUT
(3)
OUTPUT
(2)
TRIGGER
SE555/NE555
DISCHARGE
(7)
MODULATION INPUT (See Note A)
(5)
CONTROL VOLTAGE
THRESHOLD
(6)
RB
GND
(1)
C
RA = 3 kΩ
RB = 500 Ω
C = 0.01 μF
RL = 1 kΩ
See Figure 22
MODULATION INPUT VOLTAGE
OUTPUT VOLTAGE
CAPACITOR VOLTAGE
Voltage—2 V/div
Time—0.1 ms/div

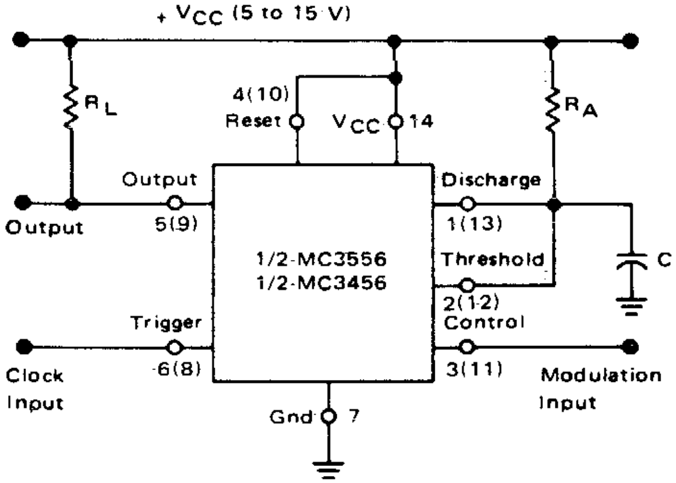
NOTE A: The modulating signal may be direct or capacitively coupled to the control voltage terminal. For direct coupling, the effects of modulation source voltage and impedance on the bias of the SE555/NE555 should be considered.

Fig. 55-3

Circuit Notes

The threshold voltage, and thereby the time delay, of a free-running oscillator is shown modulated with a triangular-wave modulation signal; however, any modulating wave-shape could be used.

PULSE-WIDTH MODULATOR



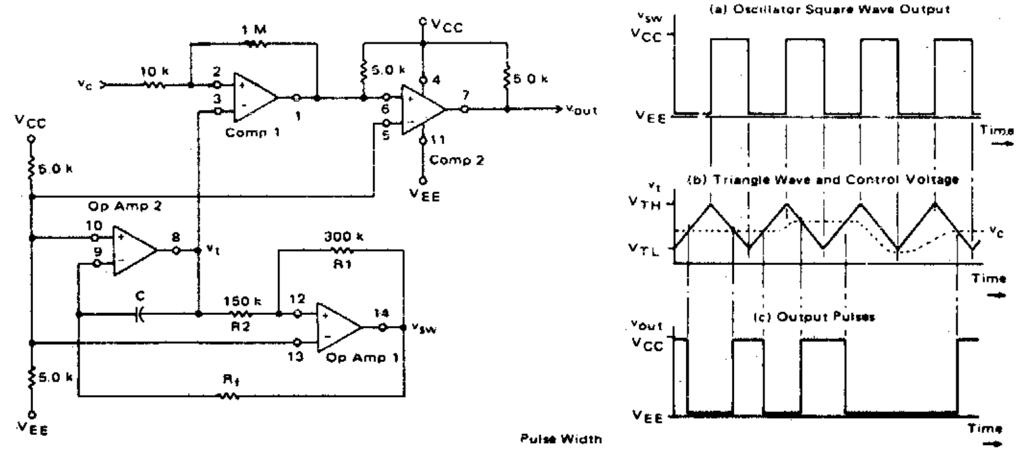
+ VCC (5 to 15 V)
RL
4(10)
Reset
VCC
14
RA
Output
Output
5(9)
Discharge
1(13)
1/2-MC3556
1/2-MC3456
Threshold
2(12)
Control
C
Trigger
Clock Input
6(8)
3(11)
Modulation Input
Gnd
7

Fig. 55-4

Circuit Notes

If the timer is triggered with a continuous pulse train in the monostable mode of operation, the charge time of the capacitor can be varied by changing the control voltage at pin 3. In this manner, the output pulse width can be modulated by applying a modulating signal that controls the threshold voltage.

PULSE-WIDTH MODULATOR


1 M
10 k
V_c
2
3
Comp 1
1
V_{CC}
5.0 k
4
6
7
5
11
Comp 2
5.0 k
V_{out}
V_{EE}
V_{CC}
5.0 k
Op Amp 2
10
9
8
V_t
300 k
R1
C
150 k
R2
12
14
V_{sw}
13
Op Amp 1
5.0 k
R_f
V_{EE}
V_{sw}
(a) Oscillator Square Wave Output
V_{CC}
V_{EE}
Time
V_t
(b) Triangle Wave and Control Voltage
V_{TH}
V_{TL}
V_c
Time
(c) Output Pulses
V_{out}
V_{CC}
V_{EE}
Time

$V_{TH} = \frac{1}{2} V_S (1 + R2/R1) + V_{EE}$ $\quad V_S = V_{CC} - V_{EE}$

$V_{TL} = \frac{1}{2} V_S (1 - R2/R1) + V_{EE}$

Oscillator Frequency

$f = \frac{R1}{4 R_f C R2}$

IC = MC3405

Pulse Width

$P.W. = \left(\frac{1}{f}\right)\left(\frac{V_c - V_{TL}}{V_{TH} - V_{TL}}\right)$ When: $V_{TL} < V_C < V_{TH}$

Duty Cycle in %

$D.C. = \left(\frac{V_c - V_{TL}}{V_{TH} - V_{TL}}\right)(100)$

Fig. 55-5

RF MODULATOR

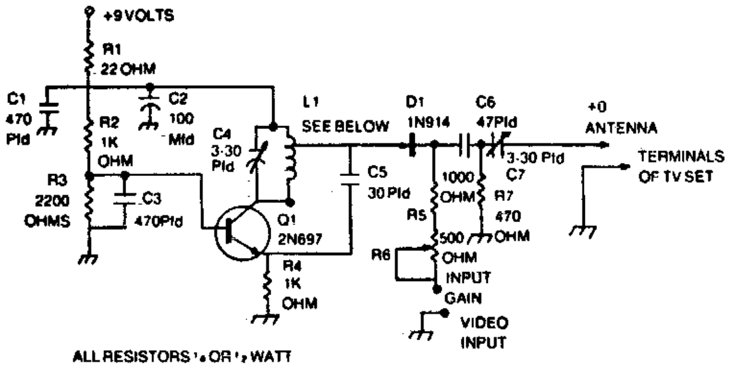
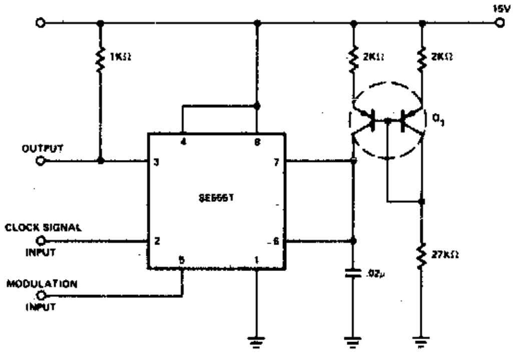

+9VOLTS
R1
22 OHM
C1
470
Pfd
C2
100
Mfd
R2
1K
OHM
C4
3-30
Pfd
L1
SEE BELOW
D1
1N914
C6
47Pfd
+0
ANTENNA
TERMINALS
OF TV SET
R3
2200
OHMS
C3
470Pfd
C5
30 Pfd
1000
OHM
R5
3-30 Pfd
C7
R7
470
OHM
Q1
2N697
R4
1K
OHM
R6
500
OHM
INPUT
GAIN
VIDEO
INPUT
ALL RESISTORS ¼ OR ½ WATT

Fig. 55-6

Circuit Notes

Capacitors C1, C3, C5, and C6 should be dipped mica. C4 and C7 are compression or piston trimmer types. R6 is PC-board mount trimpot. L1 is 6 turns of No. 14 enameled wire, ⅜ inch I.D. by ¾ inch long, tapped at 1 turn from top.

LINEAR PULSE-WIDTH MODULATOR


15V
1KΩ
2KΩ
2KΩ
Q1
OUTPUT
4
8
3
7
SE555T
CLOCK SIGNAL INPUT
2
6
5
1
MODULATION INPUT
.02μ
27KΩ

LINEARITY, ±0.2% over +4 to +12 volt input

Fig. 55-7

BALANCED MODULATOR (+12 Vdc SINGLE SUPPLY)

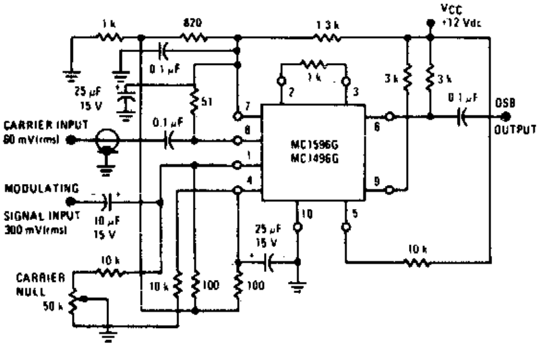
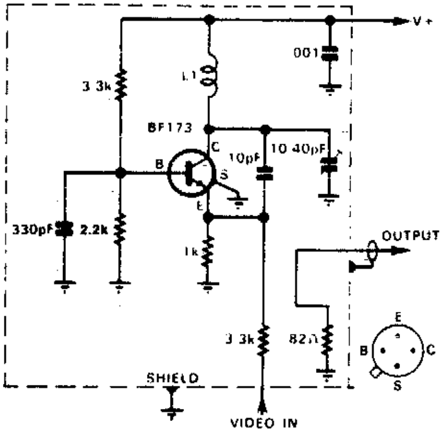

1 k
820
1.3 k
VCC +12 Vdc
0.1 μF
25 μF 15 V
1 k
2
3
3 k
3 k
51
0.1 μF
DSB OUTPUT
CARRIER INPUT 60 mV(rms)
0.1 μF
7
8
1
4
MC1596G
MC1496G
6
9
MODULATING SIGNAL INPUT 300 mV(rms)
10 μF 15 V
10
5
25 μF 15 V
10 k
CARRIER NULL 50 k
10 k
10 k
100
100

Fig. 55-8

VIDEO MODULATOR


V+
.001
3.3k
L1
BF173
C
B
S
E
10pF
10-40pF
330pF
2.2k
1k
OUTPUT
3.3k
82Ω
E
B
C
S
SHIELD
VIDEO IN

L1 10T. 22 B & S ENAMELLED COPPER WIRE. 3mm DIAMETER SEE TEXT

Fig. 55-9

MODULATOR

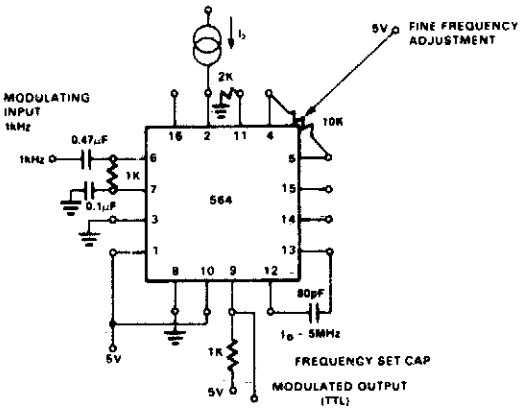
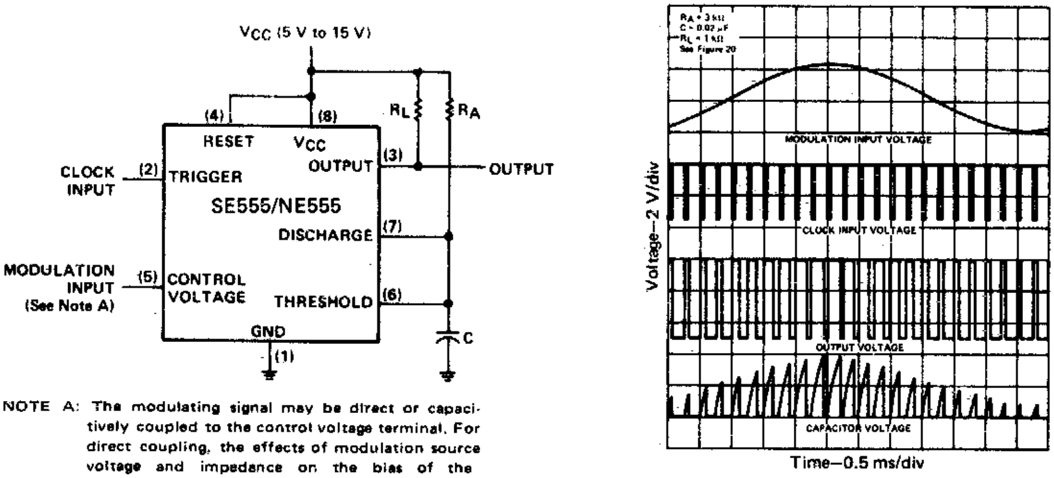

I_2
5V
FINE FREQUENCY ADJUSTMENT
2K
10K
MODULATING INPUT 1kHz
0.47μF
1kHz
1K
0.1μF
16
2
11
4
6
7
3
1
564
5
15
14
13
8
10
9
12
80pF
f_0 - 5MHz
5V
1K
5V
FREQUENCY SET CAP
MODULATED OUTPUT (TTL)

Fig. 55-10

PULSE-WIDTH MODULATOR


VCC (5 V to 15 V)
(4)
(8)
RL
RA
RESET
VCC
OUTPUT
(3)
OUTPUT
CLOCK INPUT
(2)
TRIGGER
SE555/NE555
DISCHARGE
(7)
MODULATION INPUT (See Note A)
(5)
CONTROL VOLTAGE
THRESHOLD
(6)
GND
(1)
C
RA = 3 kΩ
C = 0.02 µF
RL = 1 kΩ
See Figure 20
MODULATION INPUT VOLTAGE
CLOCK INPUT VOLTAGE
OUTPUT VOLTAGE
CAPACITOR VOLTAGE
Voltage—2 V/div
Time—0.5 ms/div

NOTE A: The modulating signal may be direct or capacitively coupled to the control voltage terminal. For direct coupling, the effects of modulation source voltage and impedance on the bias of the SE555/NE555 should be considered.

Fig. 55-11

Circuit Notes

The monostable circuit is triggered by a continuous input pulse train and the threshold voltage is modulated by a control signal. The resultant effect is a modulation of the output pulse width, as shown. A sine-wave modulation signal is illustrated, but any wave-shape could be used.

AM MODULATOR

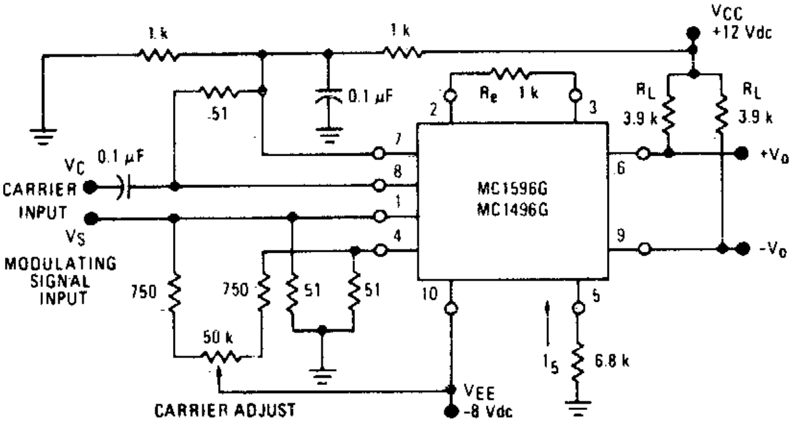
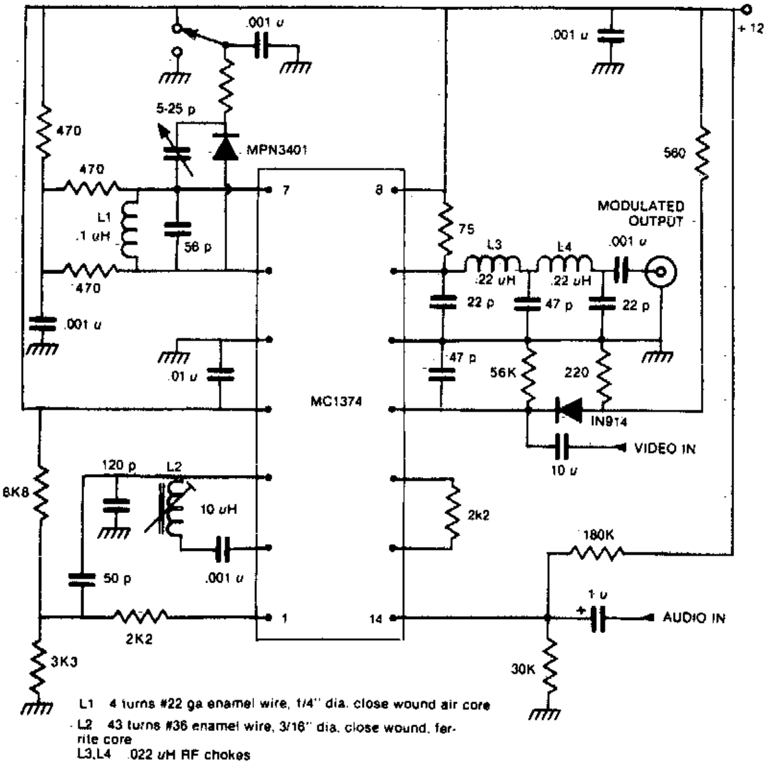

VCC +12 Vdc
1 k
1 k
.51
0.1 µF
2
Re 1 k
3
RL 3.9 k
RL 3.9 k
7
+Vo
VC 0.1 µF
CARRIER INPUT
8
6
MC1596G
MC1496G
1
VS
MODULATING SIGNAL INPUT
4
9
-Vo
750
750
51
51
10
5
50 k
I5
6.8 k
VEE -8 Vdc
CARRIER ADJUST

Fig. 55-12

TV MODULATOR USING A MOTOROLA MC1374


.001 u
+12 V
.001 u
5-25 p
MPN3401
470
470
560
L1
.1 uH
56 p
7
8
MODULATED OUTPUT
75
L3
L4
.001 u
.22 uH
.22 uH
470
22 p
47 p
22 p
.001 u
.01 u
47 p
56K
220
MC1374
IN914
VIDEO IN
10 u
120 p
L2
6K8
10 uH
2k2
180K
50 p
.001 u
1 u
1
14
AUDIO IN
2K2
3K3
30K

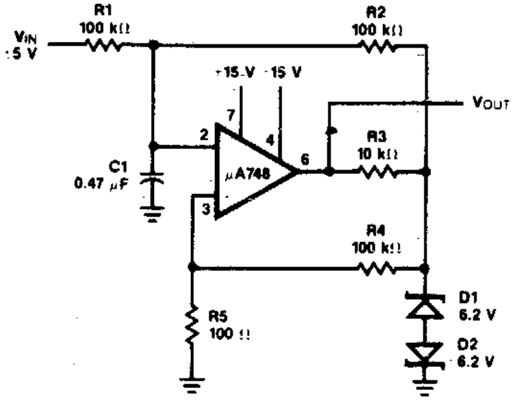
L1 4 turns #22 ga enamel wire, 1/4" dia. close wound air core
L2 43 turns #36 enamel wire, 3/16" dia. close wound, ferrite core
L3,L4 .022 uH RF chokes

Fig. 55-13

Circuit Notes

This one-chip modulator requires some outboard circuitry and a shielded box.

PULSE-WIDTH MODULATOR


R1
100 kΩ
V_{IN}
:5 V
R2
100 kΩ
+15 V
-15 V
V_{OUT}
7
2
4
6
R3
10 kΩ
C1
0.47 μF
μA748
3
R4
100 kΩ
D1
6.2 V
R5
100 Ω
D2
6.2 V

$$f_c = \frac{1}{2\pi\ R2\ C1}$$

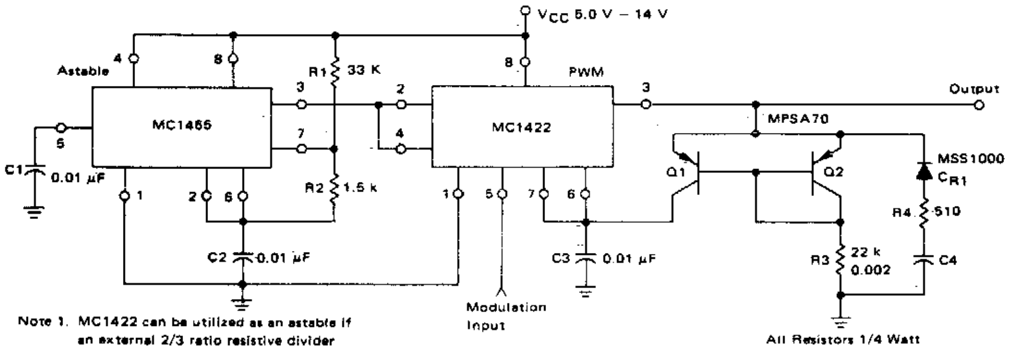
$$f_n = \frac{1}{2\pi\ R1\ C1}$$

$$= \frac{1}{2\pi\ R2\ C2}$$

$$f_c < f_n < f_{\text{unity gain}}$$

Fig. 55-14

PULSE-WIDTH MODULATOR



VCC 5.0 V – 14 V
Astable
4
8
R1
33 K
3
2
8
PWM
3
Output
MC1455
7
4
MC1422
MPSA70
5
C1
0.01 µF
1
2
6
R2
1.5 k
1
5
7
6
Q1
Q2
MSS1000
CR1
R4
510
C2
0.01 µF
C3
0.01 µF
R3
22 k
0.002
C4
Modulation Input

Note 1. MC1422 can be utilized as an astable if an external 2/3 ratio resistive divider is used at pin 5.

2. See waveforms

All Resistors 1/4 Watt

– PULSE WIDTH MODULATOR WAVEFORMS

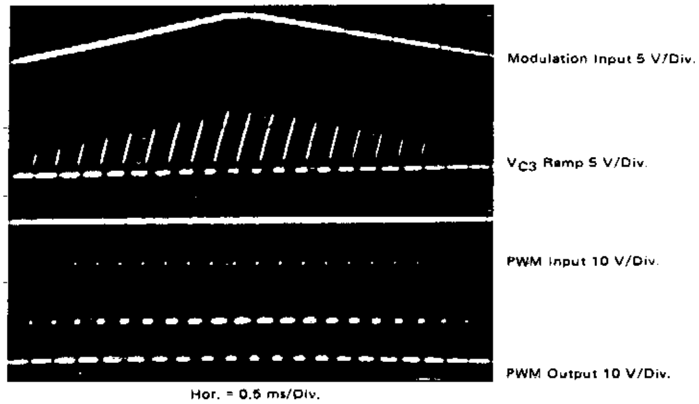
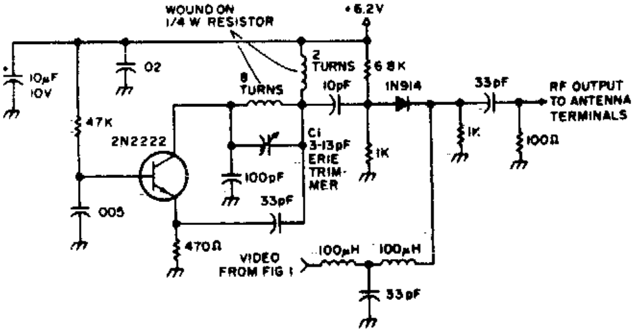


Fig. 55-15

VHF MODULATOR



WOUND ON 1/4 W RESISTOR
+6.2V
10µF 10V
.02
2 TURNS
6.8K
8 TURNS
10pF
1N914
33pF
RF OUTPUT TO ANTENNA TERMINALS
47K
2N2222
Ci 3-13pF ERIE TRIMMER
1K
1K
100Ω
100pF
33pF
.005
470Ω
VIDEO FROM FIG 1
100µH
100µH
33pF

Fig. 55-16

56

Moisture and Rain Detectors

The sources of the following circuits are contained in the Sources section beginning on page 730. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

Rain Alarm
Moisture Detector
Automatic Plant Waterer
Rain Alarm/Door Bell

RAIN ALARM

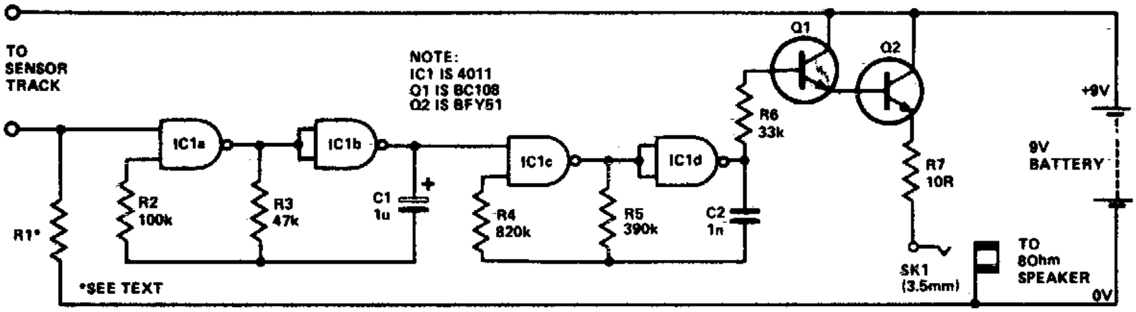
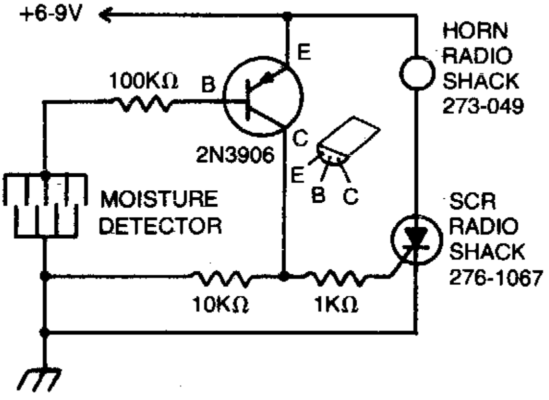

TO SENSOR TRACK
NOTE:
IC1 IS 4011
Q1 IS BC108
Q2 IS BFY51
IC1a
IC1b
IC1c
IC1d
Q1
Q2
R1*
R2 100k
R3 47k
C1 1u
+
R4 820k
R5 390k
C2 1n
R6 33k
R7 10R
+9V
9V BATTERY
SK1 (3.5mm)
TO 8Ohm SPEAKER
0V
*SEE TEXT

Fig. 56-1

Circuit Notes

The circuit uses four NAND gates of a 4011 package. In each oscillator, while one gate is configured as a straightforward inverter, the other has one input that can act as a control input. Oscillator action is inhibited if this input is held low. The first oscillator (IC1a and IC1b) has this input tied low via a high value resistor (R1) that acts as a sensitivity control. Thus this oscillator will be disabled until the control input is taken high. Any moisture bridging the sensor track will so enable the output which is a square wave at about 10 Hz. This in turn will gate on and off the 500 Hz oscillator formed by IC1c and IC1d. This latter oscillator drives the loudspeaker via R6, the Darlington pair formed by Q1 and Q2, and resistor R7.

MOISTURE DETECTOR

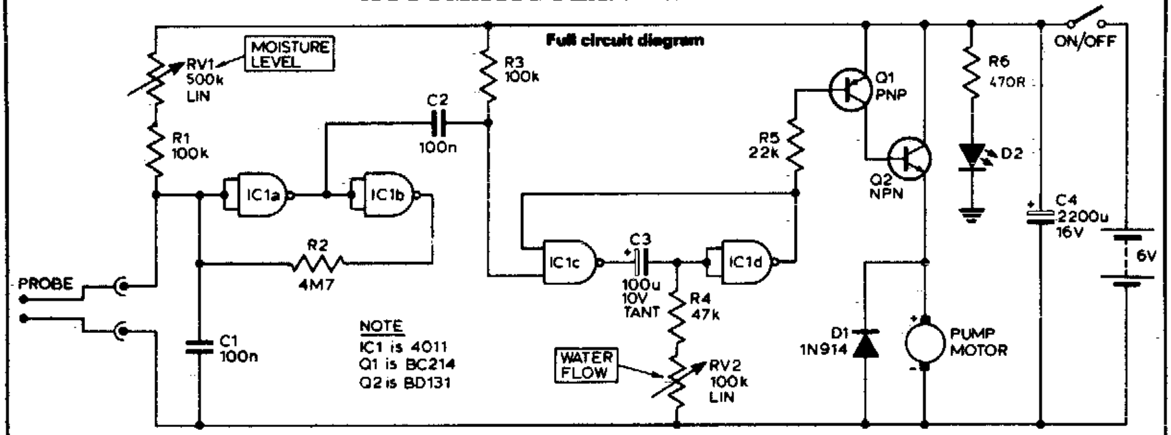

+6-9V
E
100KΩ
B
2N3906
C
E
B C
MOISTURE DETECTOR
HORN RADIO SHACK 273-049
SCR RADIO SHACK 276-1067
10KΩ
1KΩ

Circuit Notes

The detector is made of fine wires spaced about one or two inches apart. When the area between a pair of wires becomes moistened, the horn will sound. To turn it off, dc power must be disconnected.

Fig. 56-2

AUTOMATIC PLANT WATERER


Full circuit diagram
MOISTURE LEVEL
RV1 500k LIN
R1 100k
IC1a
IC1b
C2 100n
R3 100k
R2 4M7
PROBE
C1 100n
IC1c
C3 100u 10V TANT
IC1d
R4 47k
WATER FLOW
RV2 100k LIN
R5 22k
Q1 PNP
Q2 NPN
D1 1N914
PUMP MOTOR
R6 470R
D2
C4 2200u 16V
ON/OFF
6V

NOTE
IC1 is 4011
Q1 is BC214
Q2 is BD131

Fig. 56-3

Circuit Notes

The unit consists of a sensor, timer, and electric water pump. The sensor is embedded in the soil, and when dry, the electronics operate the water pump for a preset time. The circuit is composed of a level sensitive Schmitt trigger, variable time monostable, and output driver. When the resistance across the probe increases beyond a set value (i.e., the soil dries), the Schmitt is triggered. C2 feeds a negative going pulse to the monostable when the Schmitt triggers and R2 acts as feedback, to ensure a fast switching action.

RAIN ALARM/DOOR BELL

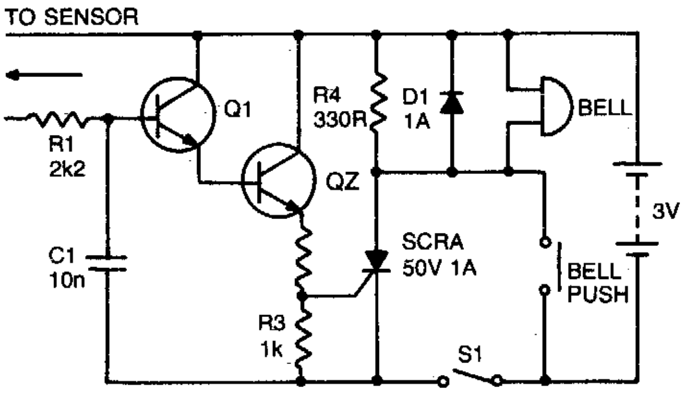

TO SENSOR
R1 2k2
C1 10n
Q1
QZ
R3 1k
R4 330R
D1 1A
SCRA 50V 1A
BELL
3V
BELL PUSH
S1

Fig. 56-4

NOTE
Q1, Q2 are 2N3706
D1 is 1N4001

Circuit Notes

With S1 open the circuit functions as a doorbell. With S1 closed, rain falling on the sensor will turn on Q1, triggering Q2 and the thyristor and activating the bell, R4 provides the holding for the thyristor while D1 prevents any damage to the thyristor from back EMF in the bell coil. The sensor can be made from 3 square inches of copper clad board with a razor cut down the center. C1 prevents any mains pickup in the sensor leads.

57

Motor Controls

The sources of the following circuits are contained in the Sources section beginning on page 730. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

Motor Speed Control
Plug-In Speed Control for Tools or Appliances
Motor Speed Control with Feedback
Direction and Speed Control for Series-Wound Motors
High-Torque Motor Speed Control
Motor Speed Control
Constant Current Motor Drive Circuit
Ac Motor Power Brake
Universal-Motor Speed Control with Load-Dependent Feedback
Dc Motor Speed/Direction Control Circuit
Servo Motor Amplifier
Motor Speed Control
Model Train Speed Control
Induction Motor Control
DC Motor Speed Control
Universal Motor Control with Built-In Self Timer
Speed Control for Model Trains or Cars
Direction and Speed Control for Shunt-Wound Motors
Two-Phase Motor Drive
Dc Servo Amplifier
Universal Motor Speed Control
Power Tool Torque Control
Ac Servo Amplifier—Bridge Type

MOTOR SPEED CONTROL

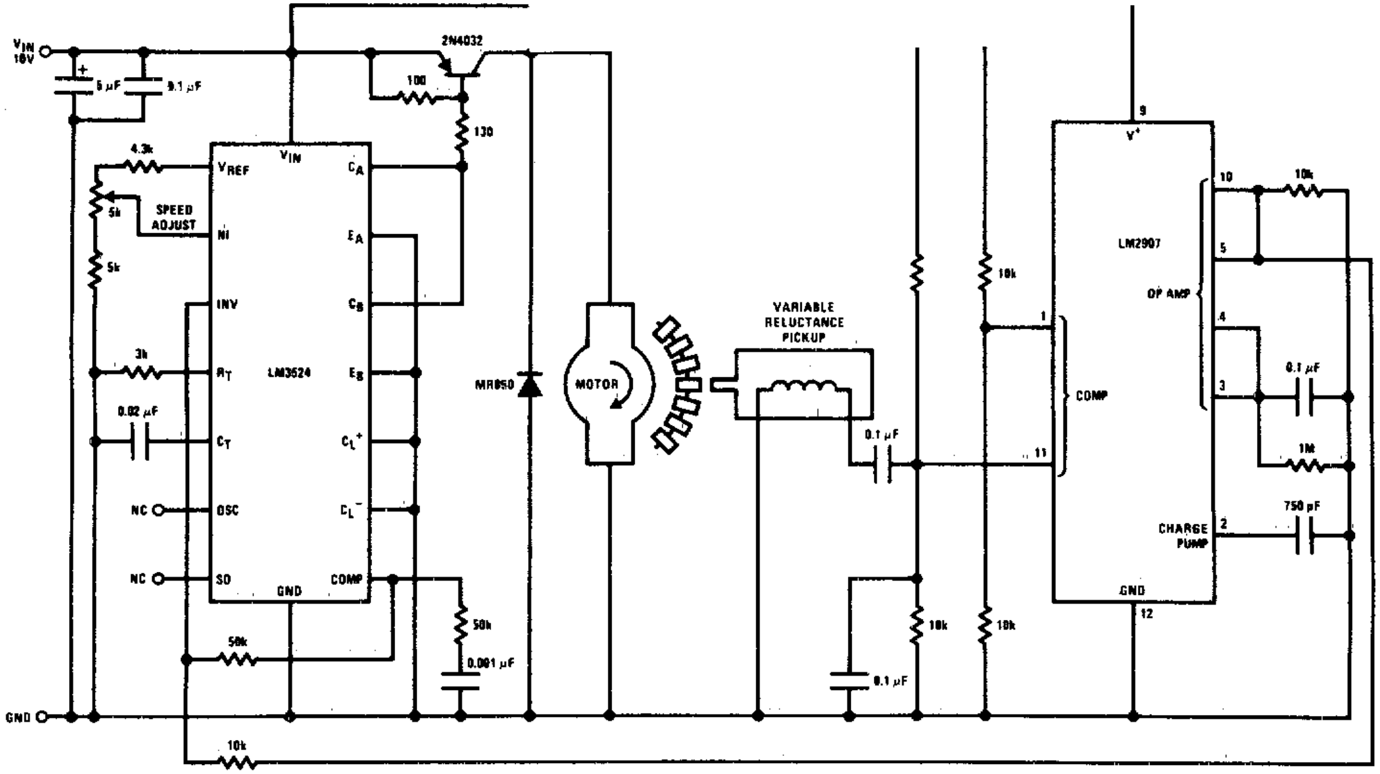
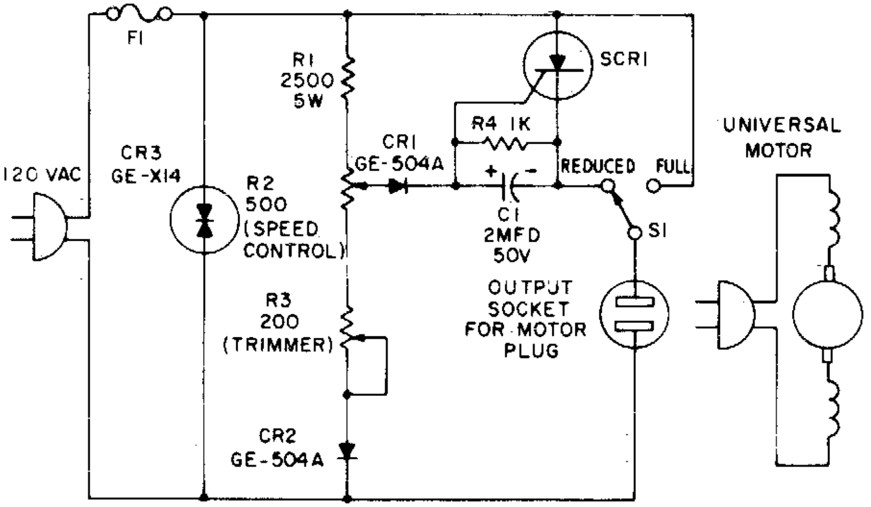

V_IN 16V
5 µF
0.1 µF
2N4032
100
130
4.3k
5k
SPEED ADJUST
5k
3k
0.02 µF
NC
NC
V_REF
V_IN
NI
INV
R_T
C_T
OSC
SD
GND
C_A
E_A
C_B
E_B
C_L+
C_L−
COMP
LM3524
50k
50k
0.001 µF
10k
MR850
MOTOR
VARIABLE RELUCTANCE PICKUP
0.1 µF
10k
10k
10k
0.1 µF
GND
9
V+
LM2907
OP AMP
COMP
CHARGE PUMP
GND
12
1
11
10
5
4
3
2
10k
0.1 µF
1M
750 pF

Fig. 57-1

Circuit Notes

This circuit is a regulating series dc motor speed control using the LM3524 for the control and drive for the motor and the LM2907 as a speed sensor for the feedback network.

PLUG-IN SPEED CONTROL FOR TOOLS OR APPLIANCES


F1
R1
2500
5W
SCR1
R4 1K
CR1
GE-504A
+
-
REDUCED
FULL
UNIVERSAL
MOTOR
120 VAC
CR3
GE-X14
R2
500
(SPEED
CONTROL)
C1
2MFD
50V
S1
OUTPUT
SOCKET
FOR MOTOR
PLUG
R3
200
(TRIMMER)
CR2
GE-504A

COMPONENT	MOTOR NAMEPLATE RATING	
	LIGHT DUTY 3 AMP MAX	HEAVIER DUTY 5 AMP MAX
SCR1	GE-X1	GE-C30B
F1	3 AMP	5 AMP

Fig. 57-2

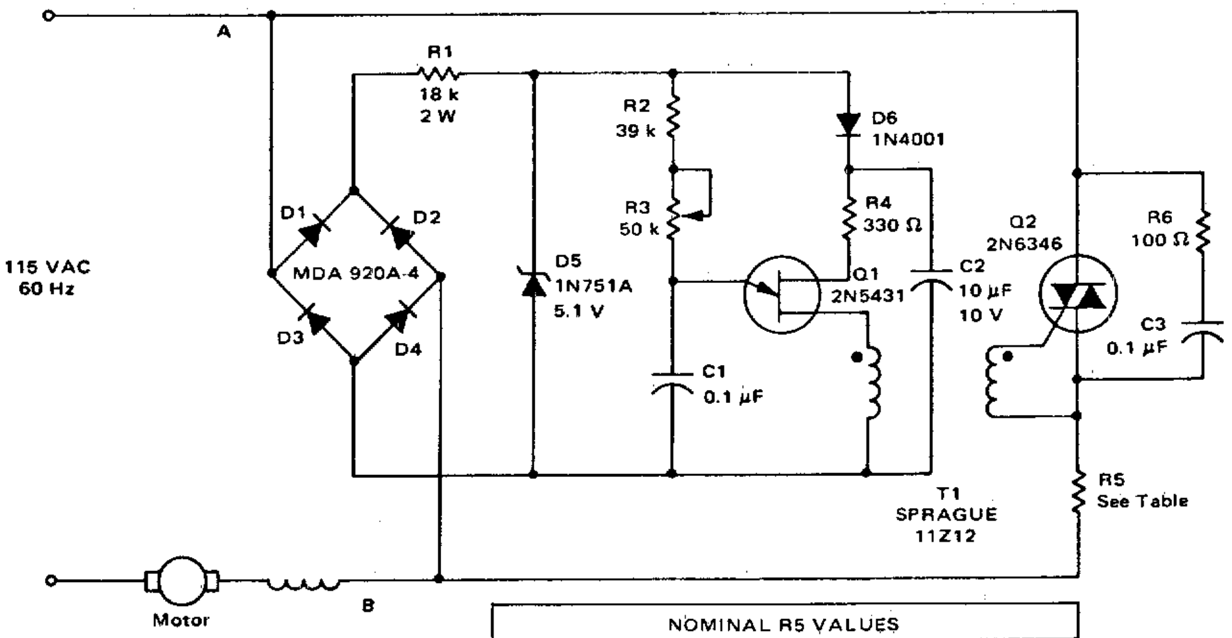
Circuit Notes

Most standard household appliances and portable hand tools can be adapted to variable-speed operation by use of this simple half-wave SCR phase control. It can be used as the speed control unit for the following typical loads providee they use series universal (brush type) motors.

Drills
Sewing Machines
Saber saws
Portable band saws
Food mixers
Food blenders
Fans
Lathes
Vibrators
Movie projectors
Sanders

During the positive half cycle of the supply voltage, the arm on potentiometer R2 taps off a traction of the sine wave supply voltage and compares it with the counter emf of the motor through the gate of the SCR. When the pot voltage rises above the armature voltage, current flows through CR1 into the gate of the SCR, triggering it, and thus applying the remainder of that half cycle supply voltage to the motor. The speed at which the motor operates can be selected by R2. Stable operation is possible over approximately a 3-to-1 speed range.

MOTOR SPEED CONTROL WITH FEEDBACK


A
B
115 VAC
60 Hz
Motor
R1
18 k
2 W
D1
D2
D3
D4
MDA 920A-4
D5
1N751A
5.1 V
R2
39 k
R3
50 k
C1
0.1 µF
D6
1N4001
R4
330 Ω
Q1
2N5431
C2
10 µF
10 V
T1
SPRAGUE
11Z12
Q2
2N6346
R6
100 Ω
C3
0.1 µF
R5
See Table

NOMINAL R5 VALUES			
Motor Rating (Amperes)	R5		
	OHMS	Watts	
2	1	5	$R5 = \frac{2}{I_M}$ I_M = Max. Rated Motor Current (RMS)
3	0.67	10	
6.5	0.32	15	

Fig. 57-3

DIRECTION AND SPEED CONTROL FOR SERIES-WOUND MOTORS

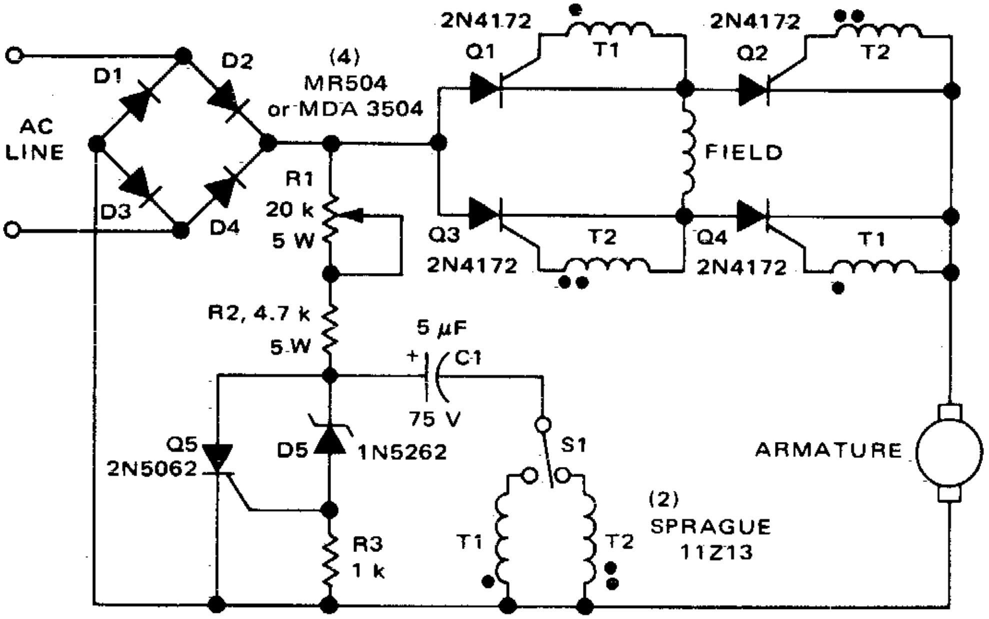
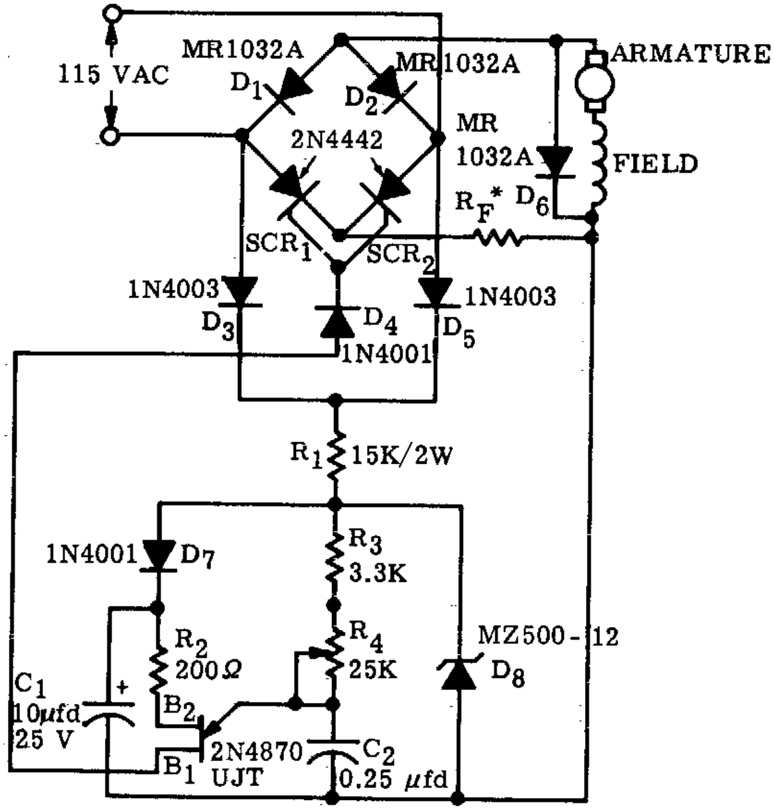

AC LINE
D1
D2
D3
D4
(4) MR504 or MDA 3504
2N4172 Q1
T1
2N4172 Q2
T2
FIELD
Q3 2N4172
Q4 2N4172
R1 20 k 5 W
R2, 4.7 k 5 W
5 μF
C1
75 V
Q5 2N5062
D5
1N5262
S1
ARMATURE
(2) SPRAGUE 11Z13
R3 1 k

Fig. 57-4

Circuit Notes

The circuit shown here can be used to control the speed and direction of rotation of a series-wound dc motor. Silicon controlled rectifiers Q1-Q4, which are connected in a bridge arrangement, are triggered in diagonal pairs. Which pair is turned on is controlled by switch S1 since it connects either coupling transformer T1 or coupling transformer T2 to a pulsing circuit. The current in the field can be reversed by selecting either SCRs Q2 and Q3 for conduction, or SCRs Q1 and Q4 for conduction. Since the armature current is always in the same direction, the field current reverses in relation to the armature current, thus reversing the direction of rotation of the motor. A pulse circuit is used to drive the SCRs through either transformer T1 or T2. The pulse required to fire the SCR is obtained from the energy stored in capacitor C1.

HIGH-TORQUE MOTOR SPEED CONTROL


115 VAC
MR1032A
D_1
MR1032A
D_2
ARMATURE
MR
1032A
D_6
FIELD
2N4442
R_F*
SCR_1
SCR_2
1N4003
D_3
D_4
1N4001
1N4003
D_5
R_1
15K/2W
1N4001
D_7
R_3
3.3K
R_2
200Ω
R_4
25K
MZ500-12
D_8
C_1
10μfd
25 V
+
B_2
B_1
2N4870
UJT
C_2
0.25 μfd

R_F* = FEEDBACK RESISTOR
(SEE TEXT)

Fig. 57-5

Circuit Notes

A bridge circuit consisting of two SCRs and two silicon rectifiers furnishes full-wave power to the motor. Diodes, D3 and D5, supply dc to the trigger circuit through dropping resistors, R1. Phase delay of SCR firing is obtained by charging C2 through resistors R3 and R4 from the voltage level established by the zener diode, D8. When C2 charges to the firing voltage of the unijunction transistor, the UJT fires, triggering the SCR that has a positive voltage on its anode. When C2 discharges sufficiently, the unijunction transistor drops out of conduction. The value of R_F is dependent upon the size of the motor and on the amount of feedback desired. A typical value for R_F can be calculated from: $R_F = \frac{2}{I_M}$. where S_{IM} is the max rated load current (rms).

MOTOR SPEED CONTROL

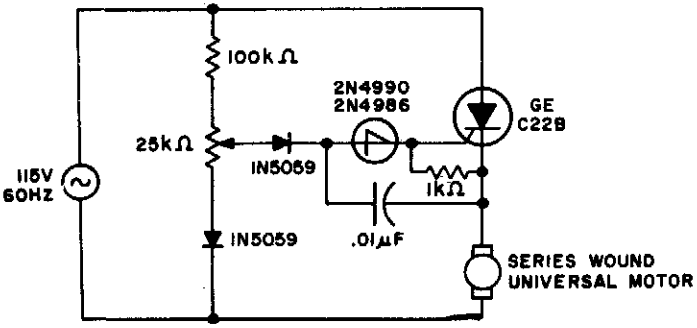

100kΩ
2N4990
2N4986
GE
C228
25kΩ
1N5059
115V
60HZ
1kΩ
1N5059
.01µF
SERIES WOUND
UNIVERSAL MOTOR

Fig. 57-6

Circuit Notes

Switching action of the 2N4990 allows smaller capacitors to be used while achieving reliable thyristor triggering.

CONSTANT CURRENT MOTOR DRIVE CIRCUIT

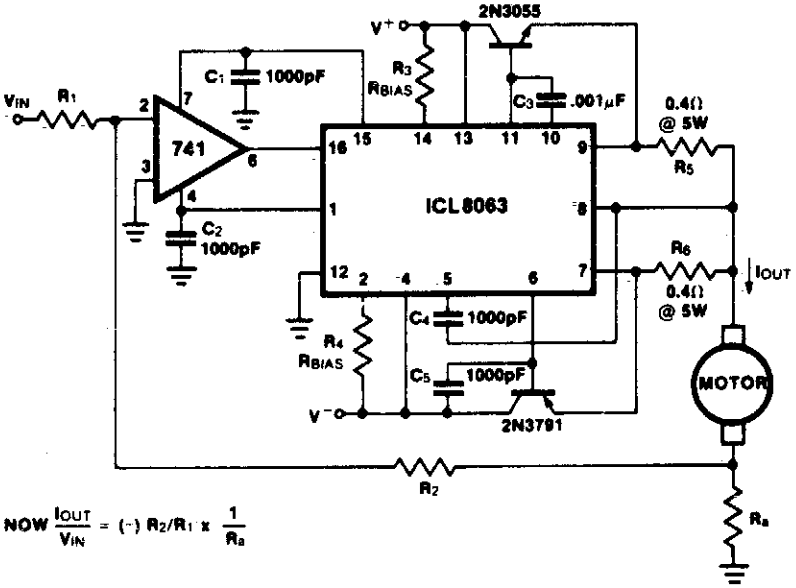
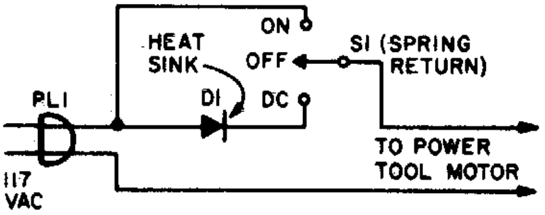

2N3055
V+
R3
RBIAS
C1
1000pF
VIN
R1
741
2
3
7
6
4
C3
.001µF
0.4Ω
@ 5W
R5
16
15
14
13
11
10
9
ICL8063
1
8
C2
1000pF
R6
12
2
4
5
6
7
0.4Ω
@ 5W
IOUT
R4
RBIAS
C4
1000pF
C5
1000pF
V−
2N3791
MOTOR
R2
Ra
NOW IOUT/VIN = (−) R2/R1 x 1/Ra

Fig. 57-7

Circuit Notes

This minimum device circuit can be used to drive dc motors where there is some likelihood of stalling or lock up; if the motor locks, the current drive remains constant and the system does not destroy itself.

AC MOTOR BRAKE


ON
HEAT SINK
OFF
SI (SPRING RETURN)
PLI
DI
DC
TO POWER TOOL MOTOR
117 VAC

PARTS LIST FOR AC MOTOR POWER BRAKE

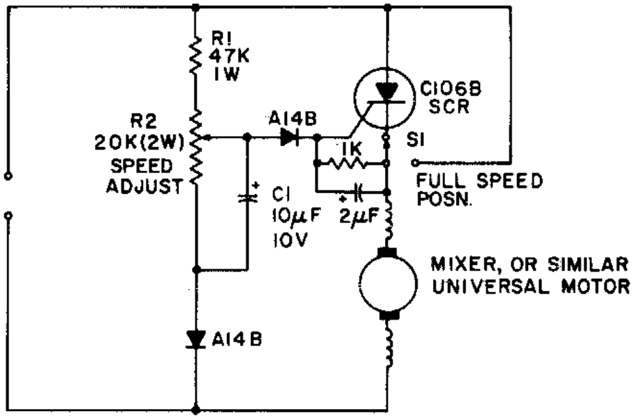
PL1—AC plug
D1—Silicon rectifier, 200 PIV, 20 A.
S1—Spdt switch. Center off, one side spring return
Misc.—Metal cabinet

Circuit Notes

A shot of direct current will instantly stop any ac power tool motor. Switch S1 is a center-off, one side spring return. With S1 on, ac will be fed to the motor and the motor will run. To brake the motor, simply press S1 down and a quick shot of dc will instantly stop it. The switch returns to the center off position when released. This Power Brake can only be used with ac motors; it will not brake universal (ac-dc) motors. A heat sink must be provided for the diode.

Fig. 57-8

UNIVERSAL-MOTOR SPEED CONTROL WITH LOAD-DEPENDENT FEEDBACK (FOR MIXER, SEWING MACHINE, ETC.)


RI 47K IW
R2 20K(2W) SPEED ADJUST
A14B
C106B SCR
IK
SI
FULL SPEED POSN.
CI 10μF 10V
2μF
MIXER, OR SIMILAR UNIVERSAL MOTOR
A14B

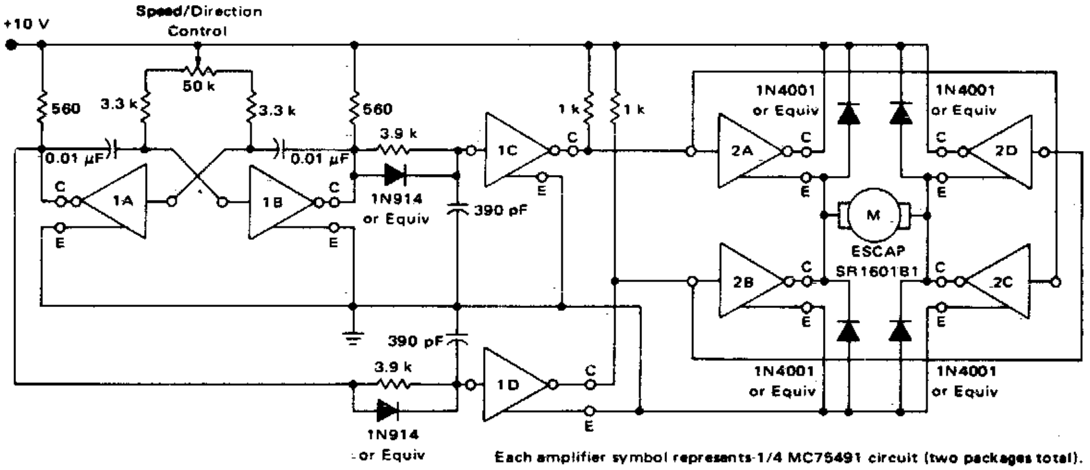
NOTE: RESISTORS 1/2 WATT EXCEPT AS NOTED

Fig. 57-9

Circuit Notes

Simple half-wave motor speed control is effective for use with small universal (ac/dc) motors. Maximum current capability 2.0 amps RMS. Because speed-dependent feedback is provided, the control gives excellent torque characteristics to the motor, even at low rotational speeds. Normal operation at maximum speed can be achieved by closing switch S1, thus bypassing the SCR.

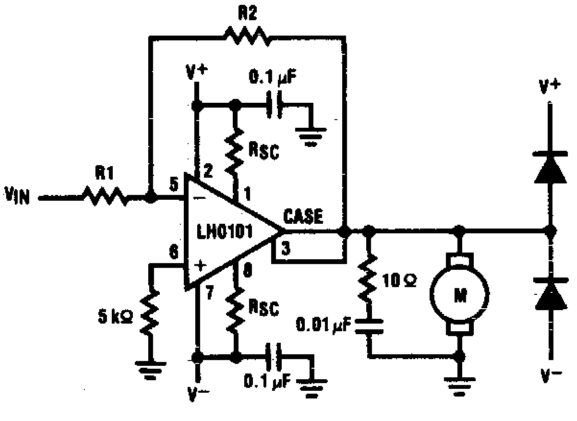
DC MOTOR SPEED/DIRECTION CONTROL CIRCUIT


+10 V
Speed/Direction Control
50 k
560
3.3 k
3.3 k
560
0.01 μF
0.01 μF
3.9 k
1N914 or Equiv
390 pF
1A
1B
1C
1D
C
E
1 k
1 k
390 pF
3.9 k
1N914 or Equiv
1N4001 or Equiv
1N4001 or Equiv
1N4001 or Equiv
1N4001 or Equiv
2A
2B
2C
2D
M
ESCAP SR1601B1

Each amplifier symbol represents 1/4 MC75491 circuit (two packages total).

Fig. 57-10

SERVO MOTOR AMPLIFIER

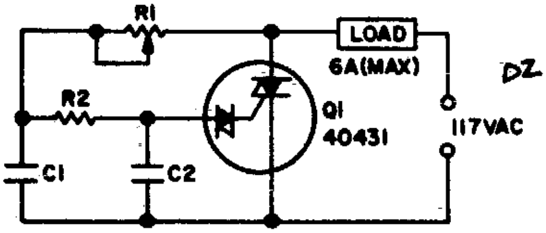

R2
V+
0.1 μF
RSC
R1
VIN
5
2
1
LH0101
CASE
3
6
8
7
5 kΩ
RSC
V−
0.1 μF
10 Ω
0.01 μF
M
V+
V−

Circuit Notes

Motor driver amplifier will deliver the rated current into the motor. Care should be taken to keep power dissipation within the permitted level. This precision speed regulation circuit employs rate feedback for constant motor current at a given input voltage.

Fig. 57-11

MOTOR SPEED CONTROL


R1
LOAD
6A(MAX)
DZ
R2
Q1
40431
117VAC
C1
C2

C1, C2—0.1-uF, 200-VDC capacitor
Q1—RCA 40431 Triac-Diac
R1—100,000-ohm linear taper potentiometer
R2—10,000-ohm, 1-watt resistor

Fig. 57-12

Circuit Notes

Universal motors and shaded-pole induction motors can be easily controlled with a full-wave Triac speed controller. Q1 combines both the triac and diac trigger diodes in the same case. The motor used for the load must be limited to 6 amperes maximum. Triac Q1 must be provided with a heat sink. With the component values shown, the Triac controls motor speed from full off to full on.

MODEL TRAIN SPEED CONTROL

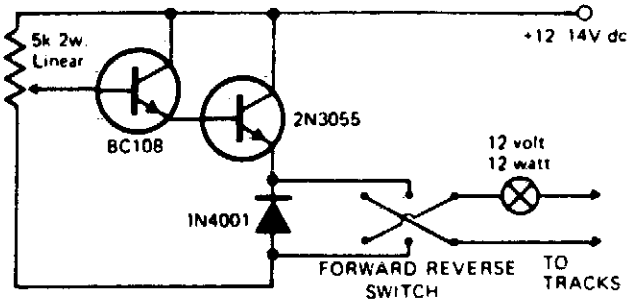
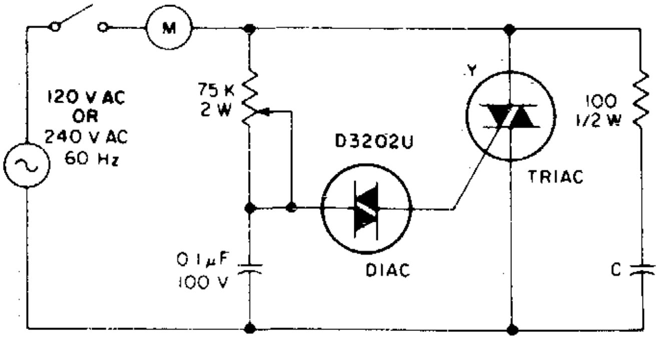

5k 2w.
Linear
+12 14V dc
BC108
2N3055
12 volt
12 watt
1N4001
FORWARD REVERSE
SWITCH
TO
TRACKS

Fig. 57-13

Circuit Notes

Virtually any NPN small signal transistor may be used in place of the BC 108 shown. Likewise any suitable NPN power transistor can be used in place of the 2N3055. The output transistor must be mounted on a suitable heatsink. Short circuit protection may be provided by wiring a 12 volt 12 watt bulb in series with the output. This will glow in event of a short circuit and thus effectively current-limit the output, it also acts as a visual short-circuit alarm.

INDUCTION-MOTOR CONTROL


M
120 V AC
OR
240 V AC
60 Hz
75 K
2 W
D3202U
0.1 μF
100 V
DIAC
Y
TRIAC
100
1/2 W
C

92CS-26324

	120VAC, 60Hz	240VAC, 60Hz
C	0.22μF, 200V	0.22μF, 400V
Y	T2700B	T2700D

Fig. 57-14

Circuit Notes

This single time-constant circuit can be used as proportional speed control for induction motors such as shaded pole or permanent split-capacitor motors when the load is fixed. The circuit is best suited to applications which require speed control in the medium to full-power range.

DC MOTOR SPEED CONTROL

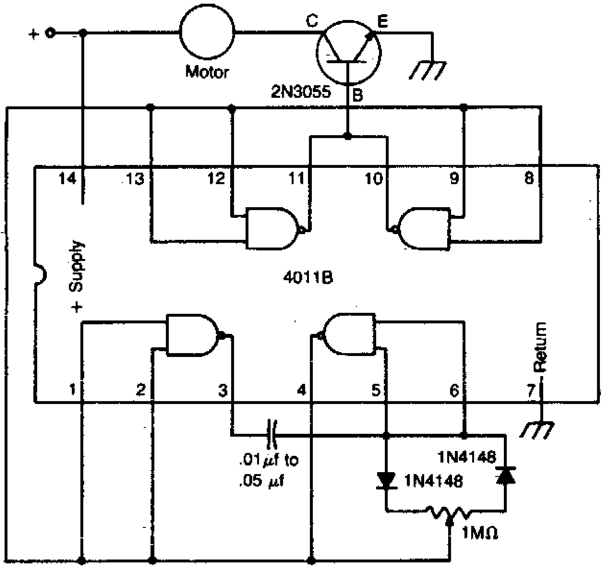

+
Motor
C
E
2N3055
B
14
13
12
11
10
9
8
+ Supply
4011B
Return
1
2
3
4
5
6
7
.01 μf to
.05 μf
1N4148
1N4148
1MΩ

Fig. 57-15

Circuit Notes

The circuit uses a 4011 CMOS NAND gate, a pair of diodes and an NPN power transistor to provide a variable duty-cycle dc source. Adjusting the speed control varies the average voltage applied to the motor. The peak voltage, however, is not changed. This pulse power is effective at very low speeds, constantly kicking the motor along. At higher speeds, the motor behaves in a nearly normal manner.

UNIVERSAL MOTOR CONTROL WITH BUILT-IN SELF TIMER

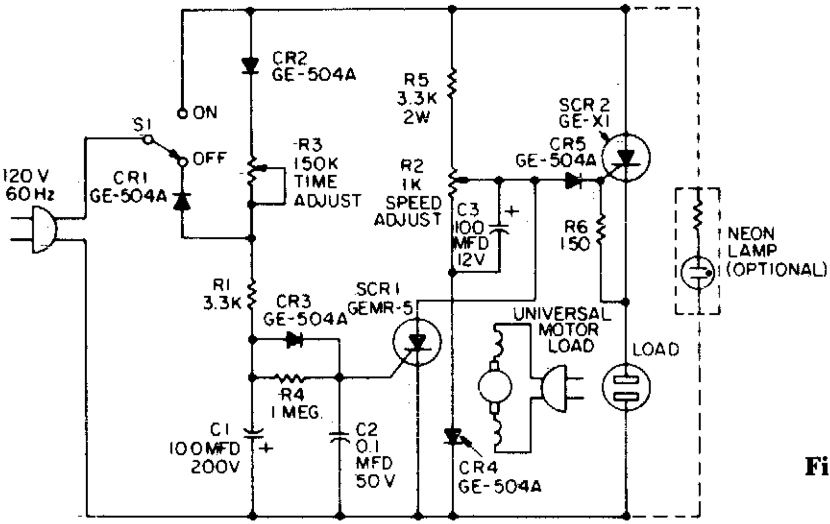

120 V
60 Hz
S1
ON
OFF
CR1
GE-504A
CR2
GE-504A
R3
150K
TIME
ADJUST
R5
3.3K
2W
R2
1K
SPEED
ADJUST
C3
100
MFD
12V
SCR 2
GE-X1
CR5
GE-504A
R6
150
NEON
LAMP
(OPTIONAL)
R1
3.3K
CR3
GE-504A
SCR 1
GEMR-5
UNIVERSAL
MOTOR
LOAD
LOAD
R4
1 MEG.
C1
100 MFD
200V
C2
0.1
MFD
50 V
CR4
GE-504A

Fig. 57-16

Circuit Notes

When the time delay expires, SCR1 conducts and removes the gate signal from SCR2, which stops the motor. Both the time delay and motor speed are adjustable by potentiometers R2 and R3. If heavier motor loads are anticipated, use the larger C30B SCR in place of the GE-X1 for SCR2. Also, the capacitance of C1 can be increased to lengthen the time delay, if desired.

SPEED CONTROL FOR MODEL TRAINS OR CARS

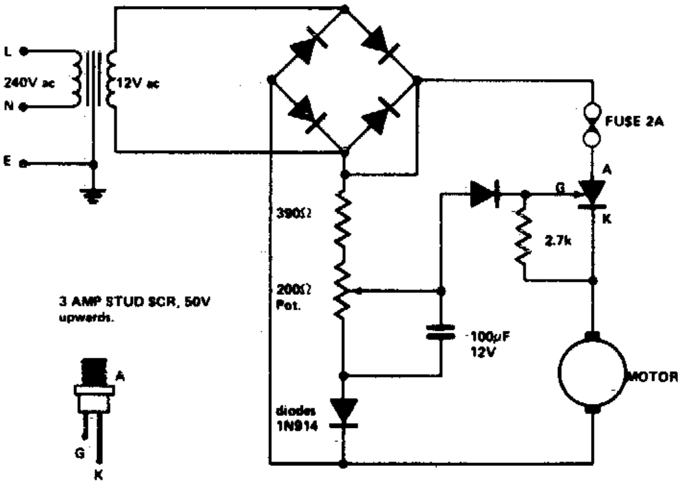
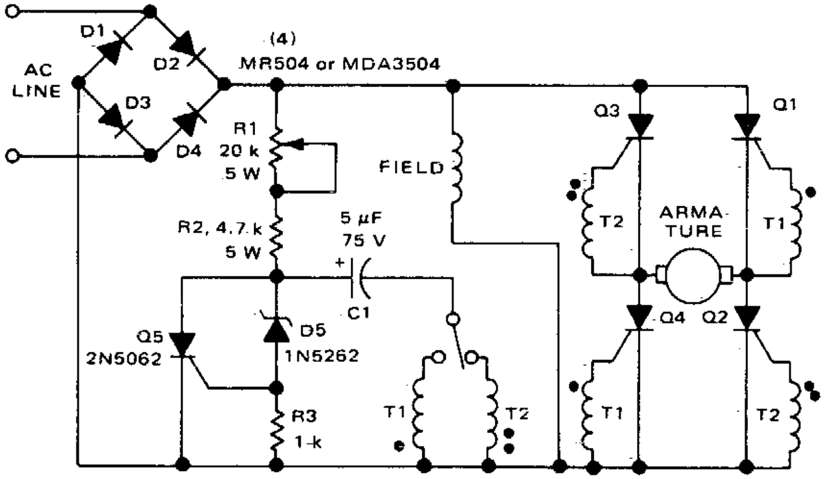

L
240V ac
N
E
12V ac
FUSE 2A
A
G
K
2.7k
390Ω
200Ω
Pot.
100μF
12V
MOTOR
diodes
1N914
3 AMP STUD SCR, 50V
upwards.
A
G
K

Fig. 57-17

Circuit Notes

Low voltage speed control gives very good starting torque and excellent speed regulation. A reversing switch may be incorporated in the leads to the motor.

DIRECTION AND SPEED CONTROL FOR SHUNT-WOUND MOTORS


AC LINE
D1
D2
D3
D4
(4)
MR504 or MDA3504
R1
20 k
5 W
R2, 4.7 k
5 W
5 µF
75 V
+
C1
FIELD
Q5
2N5062
D5
1N5262
R3
1 k
T1
T2
Q3
Q1
ARMATURE
Q4
Q2
T1
T2
T1
T2

T1 and T2 are Sprague 11Z13

Q1 thru Q4 – 2N4172

Fig. 57-18

Circuit Notes

This circuit operates like the one shown in Fig. 57-4. The only differences are that the field is placed across the rectified supply and the armature is placed in the SCR bridge. Thus the field current is unidirectional but armature current is reversible; consequently the motor's direction of rotation is reversible. Potentiometer R1 controls the speed.

TWO-PHASE MOTOR DRIVE

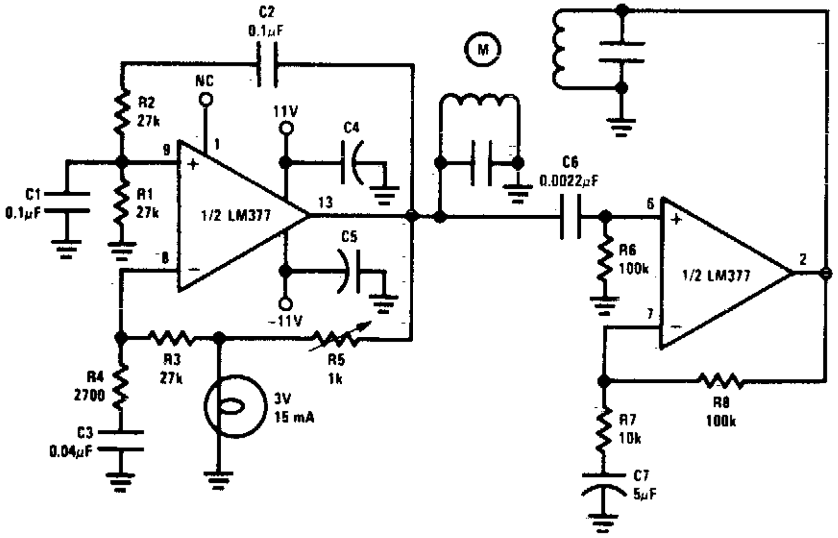

C2
0.1µF
NC
R2
27k
11V
C4
9
1
+
C1
0.1µF
R1
27k
1/2 LM377
13
C5
8
−
−11V
R4
2700
R3
27k
R5
1k
C3
0.04µF
3V
15 mA
M
C6
0.0022µF
6
+
R6
100k
1/2 LM377
2
7
−
R8
100k
R7
10k
C7
5µF

Fig. 57-19

DC SERVO AMPLIFIER

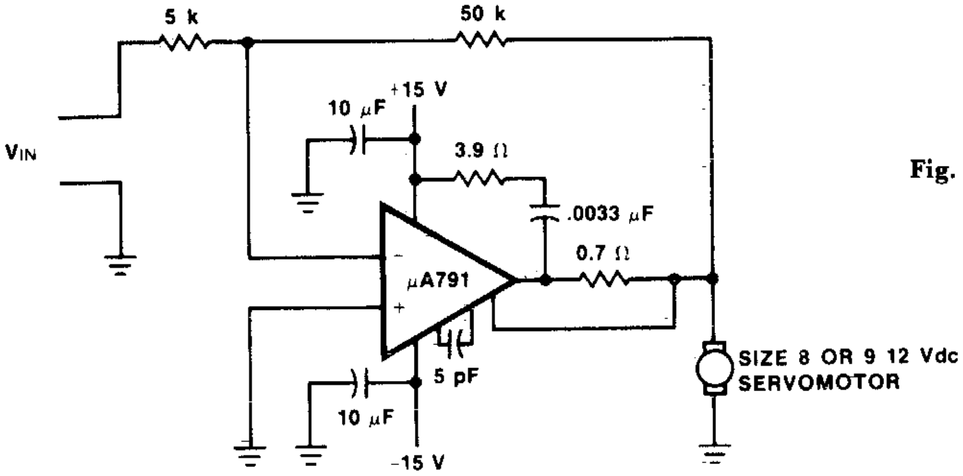
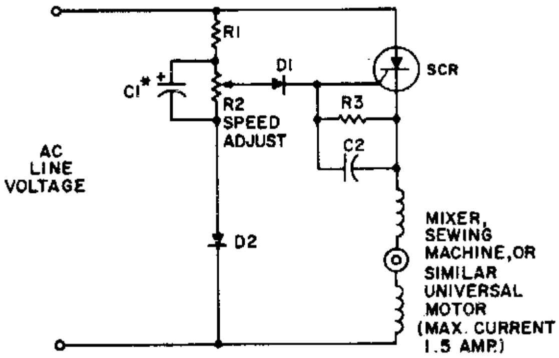

5 k
50 k
+15 V
10 μF
3.9 Ω
.0033 μF
0.7 Ω
VIN
μA741
−
+
5 pF
10 μF
−15 V
SIZE 8 OR 9 12 Vdc
SERVOMOTOR

Fig. 57-20

UNIVERSAL MOTOR SPEED CONTROL


R1
C1*
+
D1
SCR
R2
SPEED
ADJUST
R3
C2
AC
LINE
VOLTAGE
D2
MIXER,
SEWING
MACHINE,OR
SIMILAR
UNIVERSAL
MOTOR
(MAX. CURRENT
1.5 AMP.)

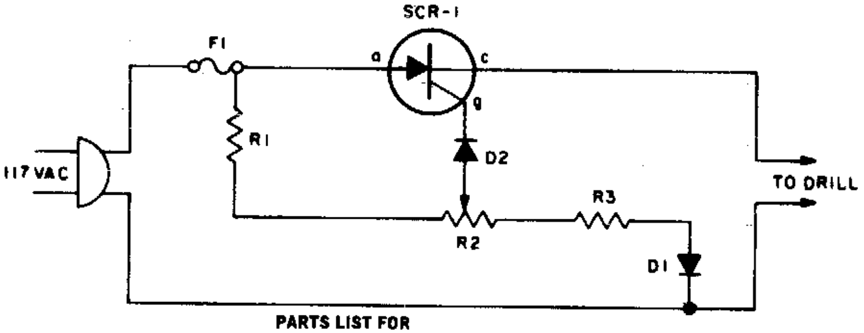
Line Voltage	120V	240V
R_1	47K	100K
R_2	10K	20K
R_3	1K	1K
C_1	1μF, 50V	1μF, 100V
C_2	0.1μF, 50V	0.1μF, 50V
D_1	1N5059	1N5060
D_2	1N5059	1N5060
SCR	C106B1	C106D1

Fig. 57-21

Circuit Notes

The resistor capacitor network R1-R2-C1 provides a ramp-type reference voltage superimposed on top of a dc voltage adjustable with the speed-setting potentiometer R2. This reference voltage appearing at the wiper of R2 is balanced against the residual counter emf of the motor through the SCR gate. As the motor slows down due to heavy loading, its counter emf falls, and the reference ramp triggers the SCR earlier in the ac cycle. More voltage is thereby applied to the motor causing it to pick up speed again. Performance with the C106 SCR is particularly good because the low trigger current requirements of this device allow use of a flat top reference voltage, which provides good feedback gain and close speed regulation.

POWER TOOL TORQUE CONTROL


SCR-1
F1
a
c
g
R1
D2
117 VAC
R3
TO DRILL
R2
D1

PARTS LIST FOR
POWER TOOL TORQUE CONTROL

D1, D2—1A, 400 PIV silicon rectifier (Calectro K4-557 or equiv.)
F1—3-A "Slo-blo" fuse
R1—2500-ohm, 5-watt resistor
R2—250-ohm, 4-watt potentiometer
R3—33-ohm, ½-watt resistor
SCR1—8-A, 400-PIV silicon controlled rectifier (HEP R1222)

Fig. 57-22

Circuit Notes

As the speed of an electric drill is decreased by loading, its torque also drops. A compensating speed control like this one puts the oomph back into the motor. When the drill slows down, a back voltage developed across the motor—in series with the SCR cathode and gate—decreases. The SCR gate voltage therefore increases relatively as the back voltage is reduced. The extra gate voltage causes the SCR to conduct over a larger angle and more current is driven into the drill, even as speed falls under load. The SCR should be mounted in ¼-in. thick block of aluminum or copper at least 1-in. square. If the circuit is used for extended periods use a 2 inch square piece.

AC SERVO AMPLIFIER—BRIDGE TYPE

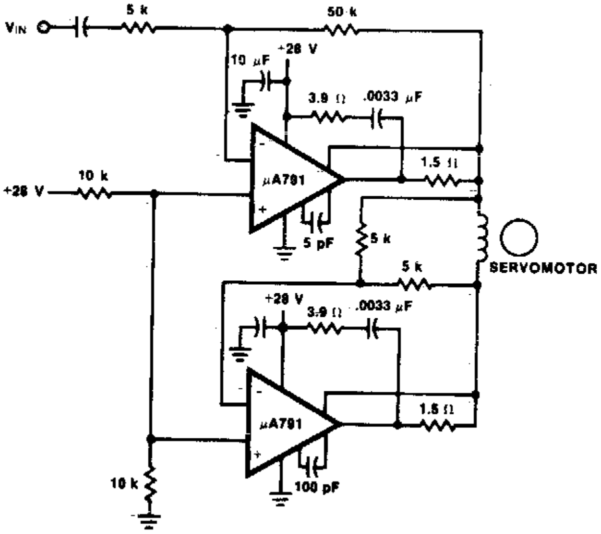

VIN
5 k
50 k
+28 V
10 µF
3.9 Ω
.0033 µF
1.5 Ω
µA791
10 k
+28 V
5 pF
5 k
5 k
SERVOMOTOR
+28 V
3.9 Ω
.0033 µF
1.5 Ω
µA791
100 pF
10 k

Fig. 57-23

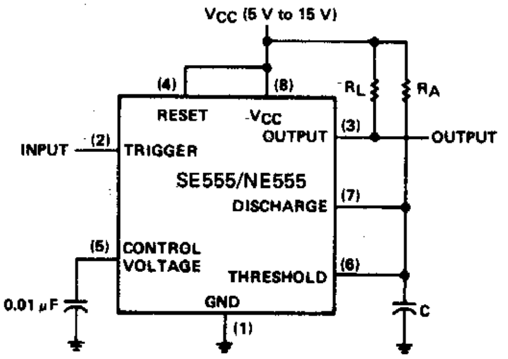
58

Multivibrators

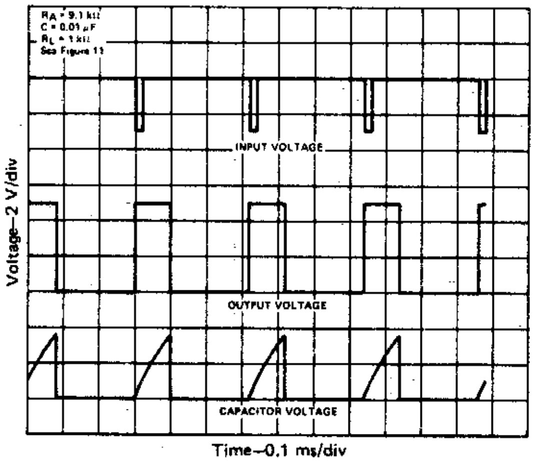
The sources of the following circuits are contained in the Sources section beginning on page 730. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

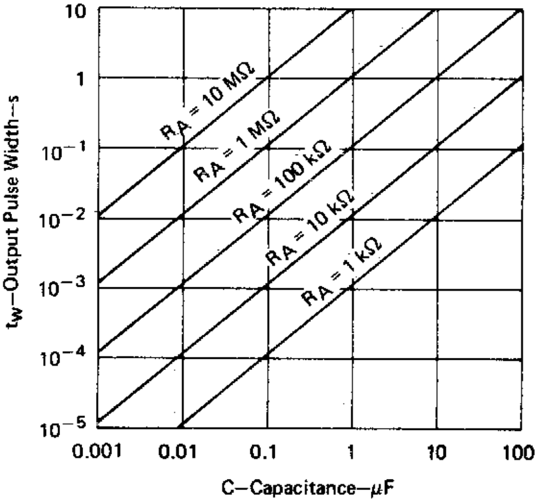
Monostable Circuit
Astable Multivibrator
Astable Oscillator
Digitally Controlled Astable Multivibrator
Dual Astable Multivibrator
UJT Monostable
Monostable Multivibrator with Input Lock-Out
TTL Monostable
Monostable Circuit
One-Shot Multivibrator
Monostable Multivibrator
Bistable Multivibrator
100 kHz Free-Running Multivibrator

MONOSTABLE CIRCUIT



VCC (5 V to 15 V)
(4)
(8)
RL
RA
RESET
VCC
OUTPUT
(3)
OUTPUT
INPUT
(2)
TRIGGER
SE555/NE555
DISCHARGE
(7)
(5)
CONTROL VOLTAGE
THRESHOLD
(6)
0.01 μF
GND
(1)
C





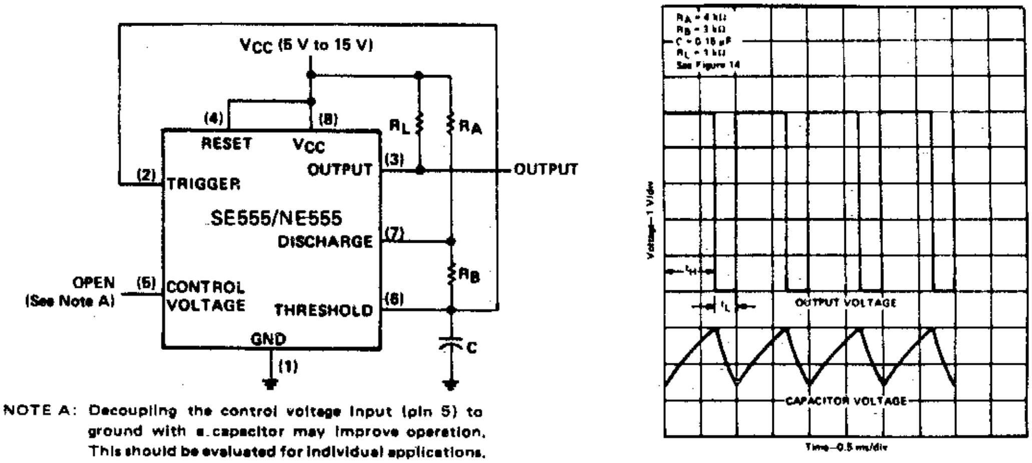
OUTPUT PULSE WIDTH vs CAPACITANCE

Fig. 58-1

Circuit Notes

If the output is low, application of a negative-going pulse to the trigger input sets the flip-flop (Q goes low), drives the output high, and turns off 1. Capacitor C is then charged through R_A until the voltage across the capacitor reaches the threshold voltage of the threshold input. If the trigger input has returned to a high level, the output of the threshold comparator will reset the flip-flop (Q goes high), drive the output low, and discharge C through Q1. Monostable operations is initiated when the trigger input voltage falls below the trigger threshold. Once initiated, the sequence will complete only if the trigger input is high at the end of the timing interval.

ASTABLE MULTIVIBRATOR


VCC (5 V to 15 V)
(4)
(8)
RESET
VCC
RL
RA
(2)
TRIGGER
OUTPUT
(3)
OUTPUT
SE555/NE555
DISCHARGE
(7)
RB
OPEN
(See Note A)
(5)
CONTROL
VOLTAGE
THRESHOLD
(6)
GND
(1)
C
RA = 4 kΩ
RB = 3 kΩ
C = 0.15 µF
RL = 1 kΩ
See Figure 14
Voltage—1 V/div
tH
tL
OUTPUT VOLTAGE
CAPACITOR VOLTAGE
Time—0.5 ms/div

NOTE A: Decoupling the control voltage input (pin 5) to ground with a capacitor may improve operation. This should be evaluated for individual applications.

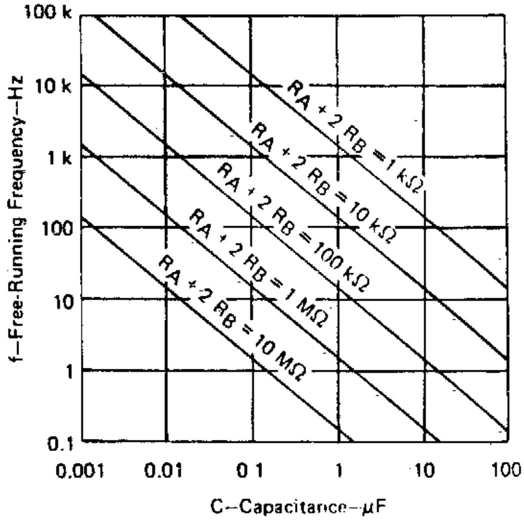


Fig. 58-2

Circuit Notes

The capacitor C will charge through R_A and R_B, and then discharge through R_B only. The duty cycle may be controlled by the values of R_A and R_B.

ASTABLE OSCILLATOR

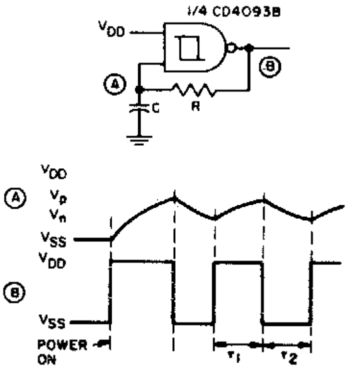
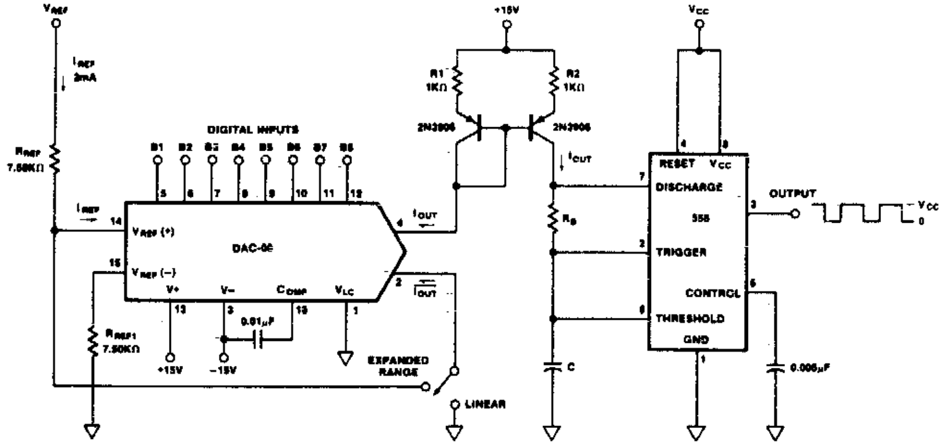

1/4 CD4093B
V_{DD}
Ⓐ
Ⓑ
C
R
V_{DD}
Ⓐ
V_p
V_n
V_{SS}
V_{DD}
Ⓑ
V_{SS}
POWER ON
T_1
T_2

Fig. 58-3

Circuit Notes

Before power is applied, the input and output are at ground potential and capacitor C is discharged. On power-on, the output goes high (V_{DD}) and C charges through R until V is reached; the output then goes low (V_{SS}). C is now discharged through R until V_n is reached. The output then goes high and charges C towards V_p through R. Thus input A alternately swings between V_p and V_n as the output goes high and low. This circuit is self-starting at power-on.

DIGITALLY CONTROLLED ASTABLE MULTIVIBRATOR

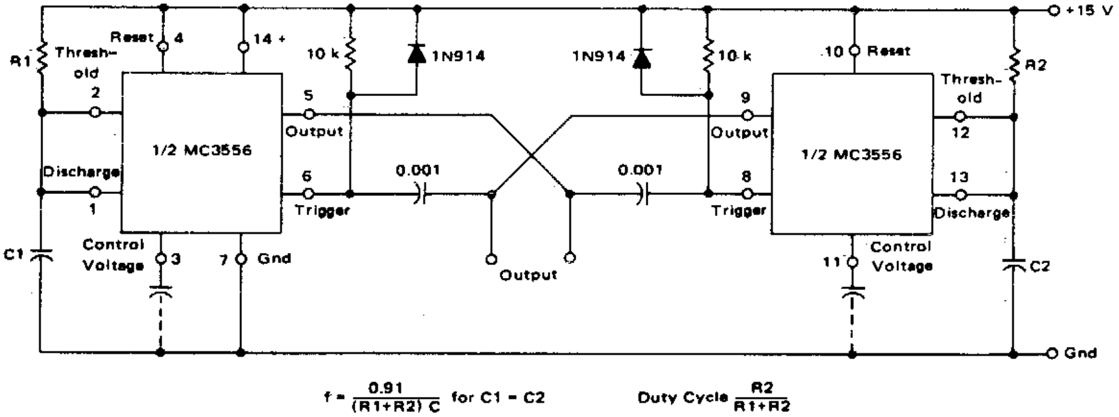

V_{REF}
I_{REF} 2mA
R_{REF} 7.50KΩ
I_{REF}
DIGITAL INPUTS
B1 B2 B3 B4 B5 B6 B7 B8
5 6 7 8 9 10 11 12
14
V_{REF} (+)
DAC-08
15
V_{REF} (−)
V+
V−
C_{OMP}
V_{LC}
13
3
16
1
0.01μF
R_{REF1} 7.50KΩ
+15V
−15V
4
I_{OUT}
2
$\overline{I_{OUT}}$
EXPANDED RANGE
LINEAR
+15V
R1 1KΩ
R2 1KΩ
2N3906
2N3906
I_{OUT}
R_B
C
V_{CC}
4
8
RESET
V_{CC}
7
DISCHARGE
555
2
TRIGGER
CONTROL
5
6
THRESHOLD
GND
1
3
OUTPUT
V_{CC}
0
0.005μF

$$\text{FREQUENCY, } f = \frac{1}{\frac{1}{3}\,\frac{R_{REF}C}{\{D\}}\,\frac{V_{CC}}{V_{REF}} + 0.695\,R_BC} \quad \text{FOR LINEAR MODE}$$

$$\text{FREQUENCY, } f = \frac{1}{\frac{1}{3}\,R_{REF}C\,\frac{V_{CC}}{V_{REF}}\left[\frac{2-\{D\}}{\{D\}}\right] + 0.695\,R_BC} \quad \text{FOR EXPANDED MODE}$$

Fig. 58-4

DUAL ASTABLE MULTIVIBRATOR



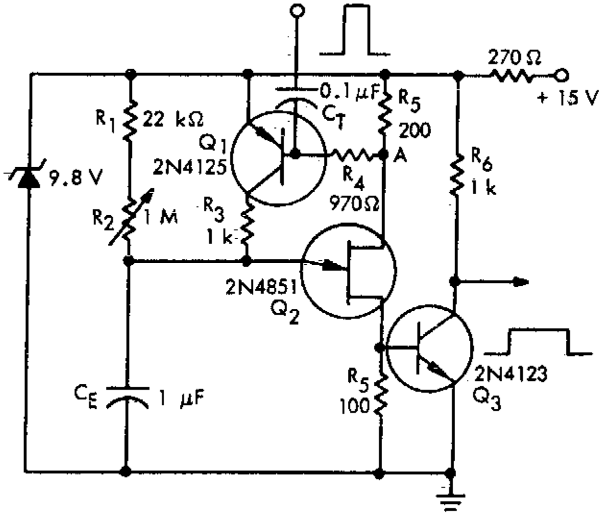
+15 V
Reset 4
14 +
R1
Thresh-old
2
1/2 MC3556
Discharge
1
C1
Control Voltage
3
7
Gnd
10 k
1N914
5
Output
6
Trigger
0.001
Output
0.001
1N914
10 k
9
Output
8
Trigger
1/2 MC3556
10 Reset
R2
Threshold
12
13
Discharge
11
Control Voltage
C2
Gnd
$f = \frac{0.91}{(R1+R2)\ C}$ for C1 = C2
Duty Cycle $\frac{R2}{R1+R2}$

Fig. 58-5

Circuit Notes

This dual astable multivibrator provides versatility not available with single timer circuits. The duty cycle can be adjusted from 5% to 95%. The two outputs provide two phase clock signals often required in digital systems. It can also be inhibited by use of either reset terminal.

UJT MONOSTABLE



270 Ω
+ 15 V
R_1 22 kΩ
Q_1
2N4125
0.1 μF
C_T
R_5
200
A
R_4
970Ω
R_6
1 k
9.8 V
R_2 1 M
R_3
1 k
2N4851
Q_2
C_E 1 μF
R_5
100
2N4123
Q_3

Fig. 58-6

MONOSTABLE MULTIVIBRATOR WITH INPUT LOCK-OUT

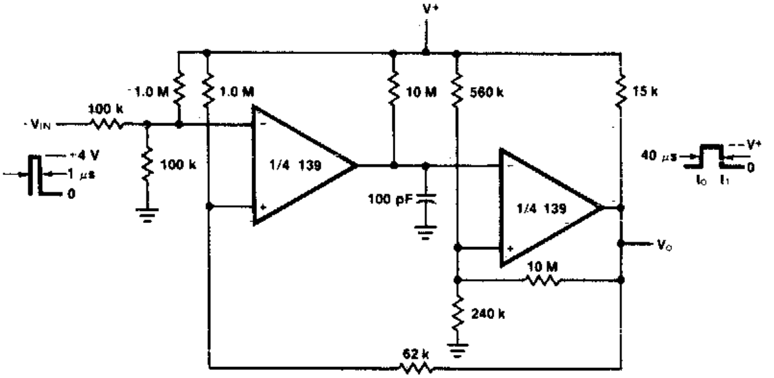

V+
1.0 M
1.0 M
10 M
560 k
15 k
100 k
- V_IN
+4 V
1 μs
0
100 k
1/4 139
100 pF
1/4 139
40 μs
V+
0
t₀ t₁
V₀
10 M
240 k
62 k

Fig. 58-7

TTL MONOSTABLE

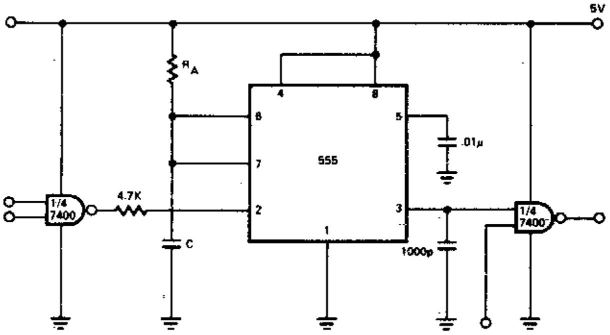

5V
R_A
4
8
6
7
555
5
.01μ
1/4 7400
4.7K
2
3
1
C
1000p
1/4 7400

Fig. 58-8

MONOSTABLE CIRCUIT

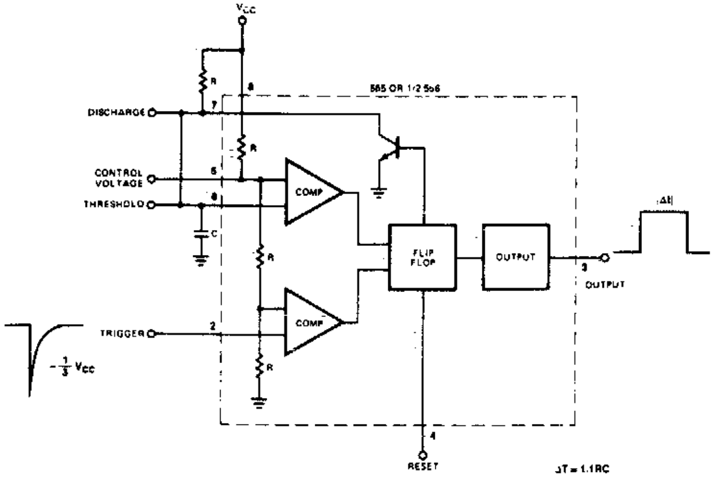

V_CC
R
8
555 OR 1/2 556
DISCHARGE
7
R
CONTROL VOLTAGE
5
THRESHOLD
6
C
COMP
FLIP FLOP
OUTPUT
ΔI
3
OUTPUT
R
COMP
TRIGGER
2
−1/3 Vcc
R
4
RESET
ΔT = 1.1RC

Fig. 58-9

ONE-SHOT MULTIVIBRATOR

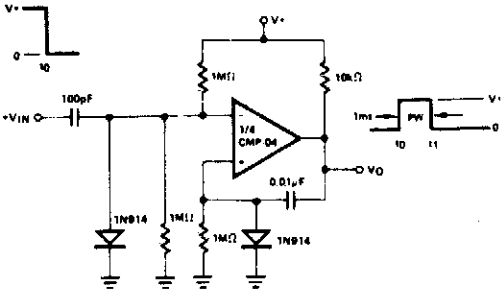

V+
0
t0
O V+
1MΩ
10kΩ
100pF
+VIN
1/4
CMP-04
1ms
PW
V+
0
t0
t1
VO
0.01μF
1N914
1MΩ
1MΩ
1N914

Fig. 58-10

MONOSTABLE MULTIVIBRATOR

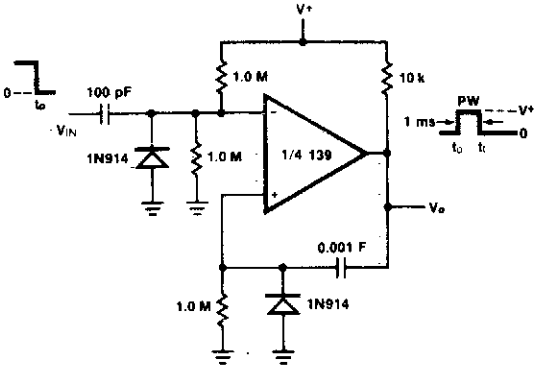

V+
0
t0
100 pF
VIN
1.0 M
10 k
PW
V+
1 ms
0
t0
t1
1N914
1.0 M
1/4 139
Vo
0.001 F
1.0 M
1N914

Fig. 58-11

BISTABLE MULTIVIBRATOR

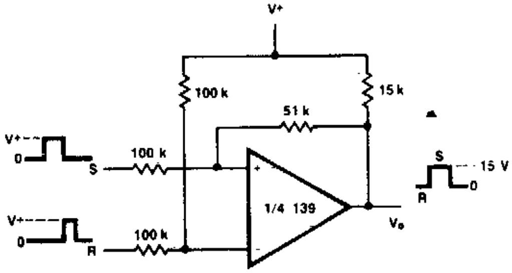
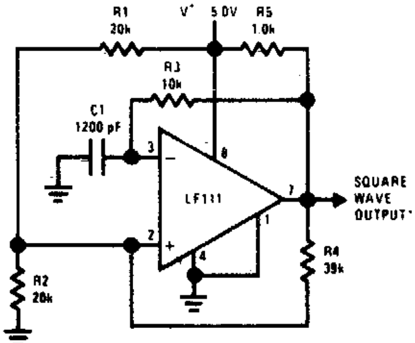

V+
100 k
15 k
51 k
V+
0
S
100 k
S
15 V
0
R
1/4 139
V+
0
R
100 k
Vo

Fig. 58-12

100 kHz FREE-RUNNING MULTIVIBRATOR


R1
20k
V+
5.0V
R5
1.0k
R3
10k
C1
1200 pF
3
8
LF111
7
SQUARE
WAVE
OUTPUT*
1
2
4
R2
20k
R4
39k

*TTL or DTL fanout of two.

Fig. 58-13

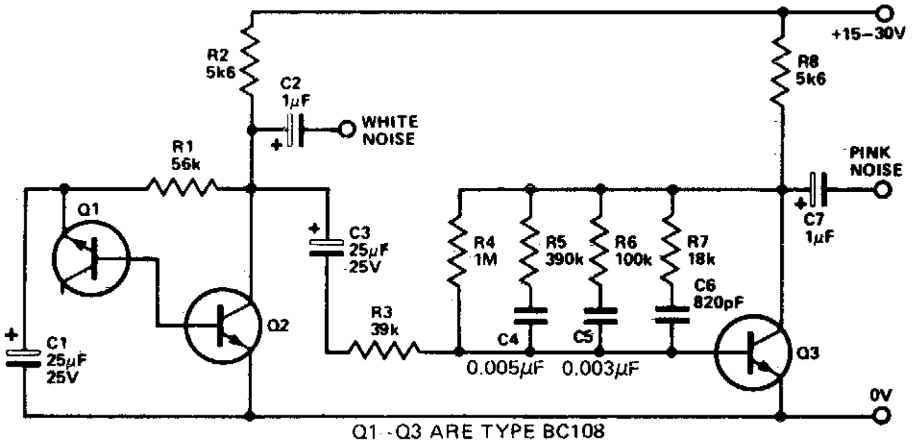
59

Noise Generators

The sources of the following circuits are contained in the Sources section beginning on page 730. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

Audio Noise Generator
Pink Noise Generator
Noise Generator
Wideband Noise Generator
Noise Generator Circuit

AUDIO NOISE GENERATOR



+15–30V
R2
5k6
C2
1µF
WHITE NOISE
R1
56k
Q1
C3
25µF
25V
R3
39k
Q2
+
C1
25µF
25V
R4
1M
R5
390k
R6
100k
R7
18k
C6
820pF
C4
C5
0.005µF
0.003µF
R8
5k6
PINK NOISE
C7
1µF
Q3
0V
Q1 - Q3 ARE TYPE BC108

Fig. 59-1

Circuit Notes

This simple circuit generates both white and pin noise. Transistor Q1 is used as a zener diode. The normal base-emitter junction is reverse-biased and goes into zener breakdown at about 7 to 8 volts. The zener noise current from Q1 flows into the base of Q2 such that an output of about 150 millivolts of white noise is available. To convert the white noise to pink, a filter is required which provides a 3 dB cut per octave as the frequency increases. Since such a filter attenuates the noise considerably an amplifier is used to restore the output level. Transistor Q3 is this amplifier and the pink noise filter is connected as a feedback network. between collector and base in order to obtain the required characteristic by controlling the gain-versus-frequency of the transistor. The output of transistor Q3 is thus the pink noise required and is fed to the relevant output socket.

PINK NOISE GENERATOR

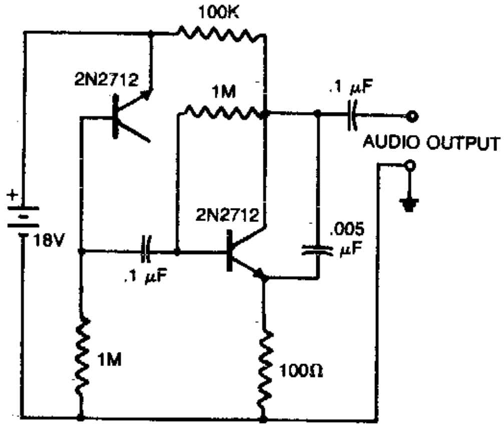

100K
2N2712
1M
.1 μF
AUDIO OUTPUT
+
18V
2N2712
.005 μF
.1 μF
1M
100Ω

Fig. 59-2

Circuit Notes

A reverse-biased pn junction of a 2N2712 transistor is used as a noise generator. The second 2N2712 is an audio amplifier. The 0.005 μF capacitor across the amplifier output removes some high-frequency components to simulate pink noise more closely. The audio output may be connected to high-impedance earphones or to a driver amplifier for speaker listening.

NOISE GENERATOR

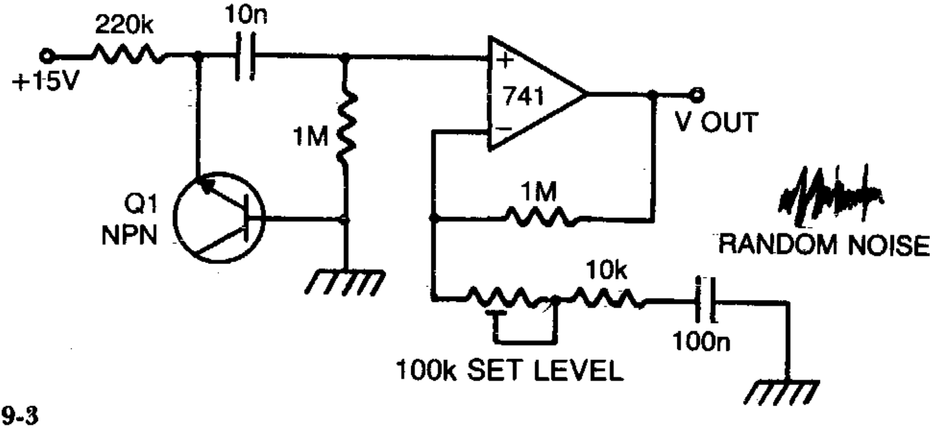

220k
10n
+15V
741
+
−
V OUT
1M
Q1
NPN
1M
RANDOM NOISE
10k
100n
100k SET LEVEL

Fig. 59-3

Circuit Notes

The zener breakdown of a transistor junction is used as a noise generator. The breakdown mechanism is random and this voltage has a high source impedance. By using the op amp as a high input impedance, high ac gain amplifier, a low impedance, large signal noise source is obtained. The 100K potentiometer is used to set the noise level by varying the gain from 40 to 20 dB.

WIDEBAND NOISE GENERATOR

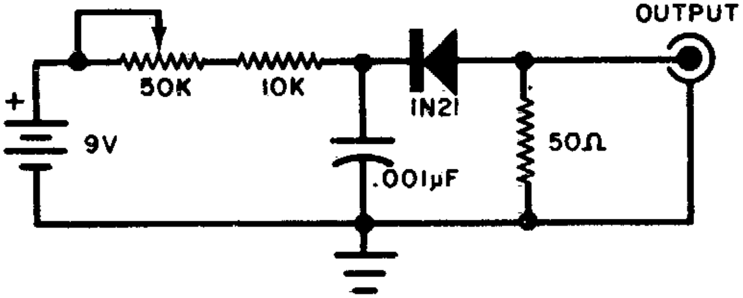

OUTPUT
+
9V
50K
10K
1N21
.001µF
50Ω

Fig. 59-4

Circuit Notes

This circuit will produce wideband rf noise. It uses a reverse-biased diode and has a low-impedance output. Can be used to align receivers for optimum performance.

NOISE GENERATOR CIRCUIT

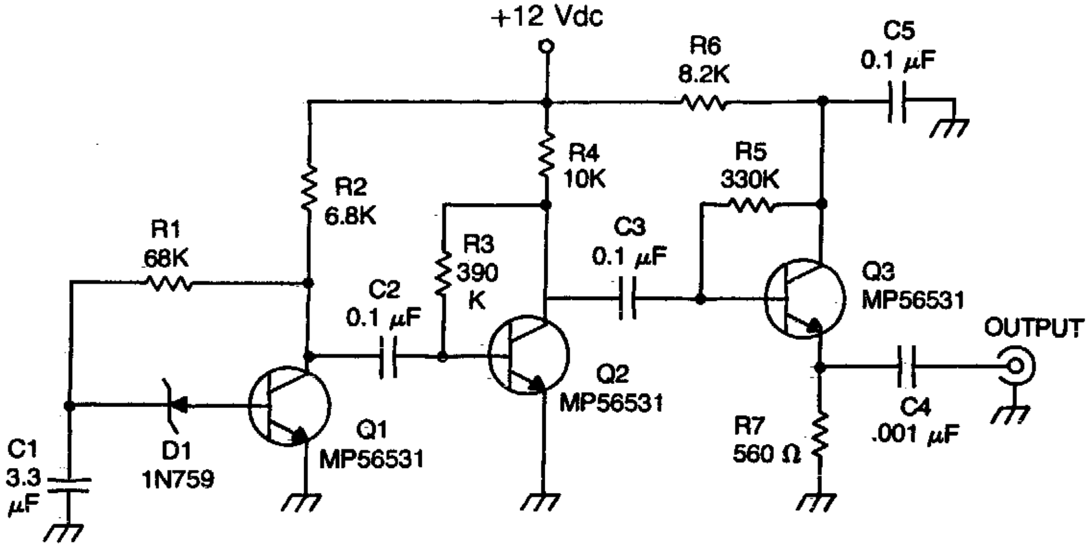

+12 Vdc
R6
8.2K
C5
0.1 µF
R4
10K
R5
330K
R2
6.8K
R1
68K
R3
390
K
C3
0.1 µF
C2
0.1 µF
Q3
MP56531
OUTPUT
Q2
MP56531
Q1
MP56531
C4
.001 µF
R7
560 Ω
C1
3.3
µF
D1
1N759

Fig. 59-5

Circuit Notes

The zener diode is an avalanche rectifier in the reverse bias mode connected to the input circuit of a wideband rf amplifier. The noise is amplified and applied to the cascade wideband amplifier, transistors Q2 and Q3.

60

Oscilloscope Circuits

The sources of the following circuits are contained in the Sources section beginning on page 730. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

Oscilloscope Converter Provides Four-Channel Displays
Add-On Triggered Sweep
10.7 MHz Sweep Generator
Drawing Circles on a Scope
Transmitter-Oscilloscope Coupler for CB Signals
Oscilloscope Monitor
Beam Splitter for Oscilloscope

OSCILLOSCOPE CONVERTER PROVIDES FOUR-CHANNEL DISPLAYS

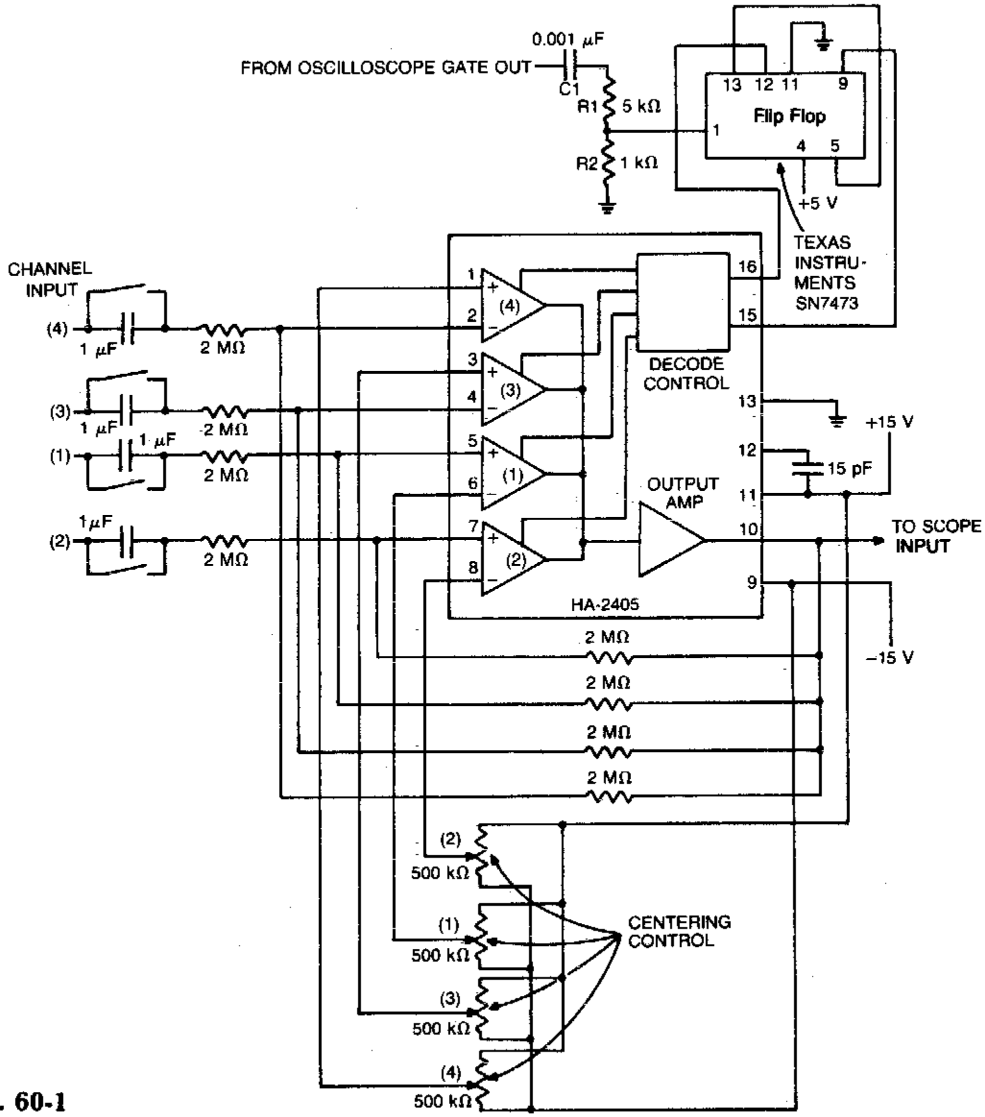

0.001 μF
FROM OSCILLOSCOPE GATE OUT
C1
R1
5 kΩ
R2
1 kΩ
13 12 11 9
Flip Flop
1
4 5
+5 V
TEXAS INSTRUMENTS SN7473
CHANNEL INPUT
(4)
1 μF
2 MΩ
(3)
1 μF
2 MΩ
(1)
1 μF
2 MΩ
(2)
1μF
2 MΩ
(4)
(3)
(1)
(2)
DECODE CONTROL
OUTPUT AMP
16
15
13
12
11
10
9
+15 V
15 pF
TO SCOPE INPUT
−15 V
HA-2405
2 MΩ
2 MΩ
2 MΩ
2 MΩ
500 kΩ
500 kΩ
500 kΩ
500 kΩ
CENTERING CONTROL

Fig. 60-1

Circuit Notes

The monolithic quad operational amplifier provides an inexpensive way to increase display capability of a standard oscilloscope. Binary inputs drive the IC op amp; a dual flip-flop divides the scope's gate output to obtain channel selection signals. All channels have centering controls for nulling offset voltage. A negative-going scope gate signal selects the next channel after each trace. The circuit operates out to 5 MHz.

ADD-ON TRIGGERED SWEEP

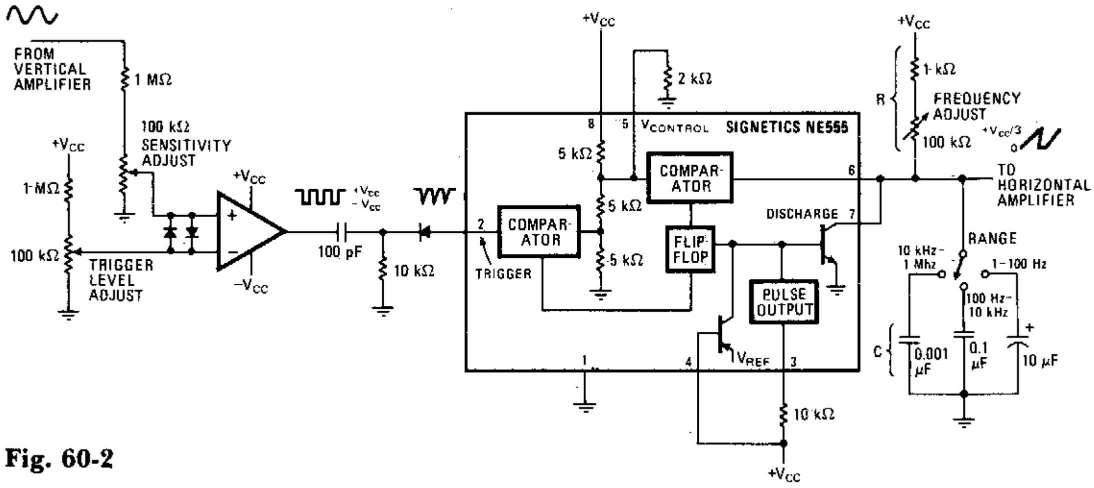
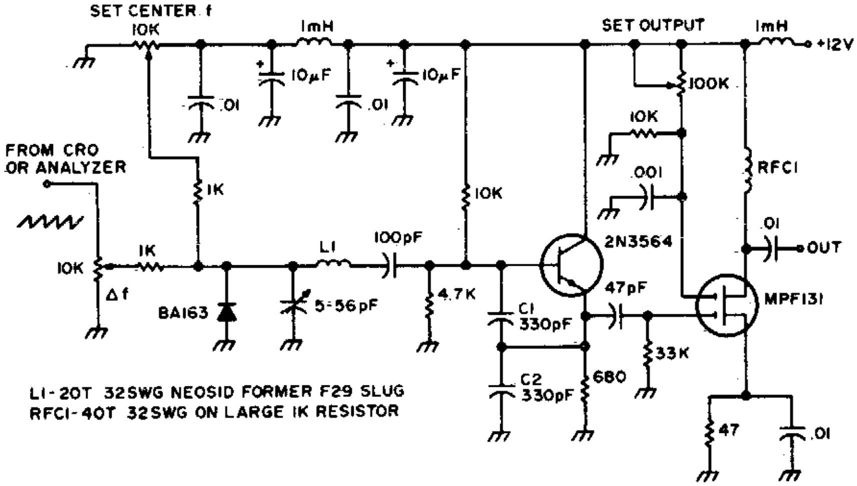

FROM VERTICAL AMPLIFIER
1 MΩ
100 kΩ SENSITIVITY ADJUST
+Vcc
1 MΩ
100 kΩ
TRIGGER LEVEL ADJUST
+Vcc
−Vcc
+Vcc
−Vcc
100 pF
10 kΩ
+Vcc
2 kΩ
8
5
VCONTROL
SIGNETICS NE555
5 kΩ
COMPARATOR
5 kΩ
2
COMPARATOR
TRIGGER
5 kΩ
FLIP FLOP
DISCHARGE
7
6
PULSE OUTPUT
1
4
VREF
3
10 kΩ
+Vcc
+Vcc
R
1 kΩ
FREQUENCY ADJUST
100 kΩ
+Vcc/3
0
TO HORIZONTAL AMPLIFIER
RANGE
10 kHz-1 Mhz
1-100 Hz
100 Hz-10 kHz
C
0.001 µF
0.1 µF
10 µF
+

Fig. 60-2

Circuit Notes

The circuit's input op amp triggers the timer, setting its flip-flop and cutting off its discharge transistor so that capacitor C can charge. When the capacitor voltage reaches the timer's control voltage (0.33Vcc), the flip-flop resets and the transistor conducts, discharging the capacitor.

10.7 MHz SWEEP GENERATOR


SET CENTER f
10K
1mH
SET OUTPUT
1mH
+12V
10µF
.01
10µF
.01
100K
10K
.001
RFC1
FROM CRO OR ANALYZER
1K
10K
.01
OUT
10K
1K
L1
100pF
2N3564
Δf
BA163
5-56pF
4.7K
C1 330pF
47pF
MPF131
33K
C2 330pF
680
47
.01

L1-20T 32SWG NEOSID FORMER F29 SLUG
RFC1-40T 32SWG ON LARGE 1K RESISTOR

Fig. 60-3

Circuit Notes

This circuit is used to observe the response of an if amp or a filter. It can be used with an oscilloscope or, for more dynamic range, with a spectrum analyzer.

DRAWING CIRCLES ON A SCOPE

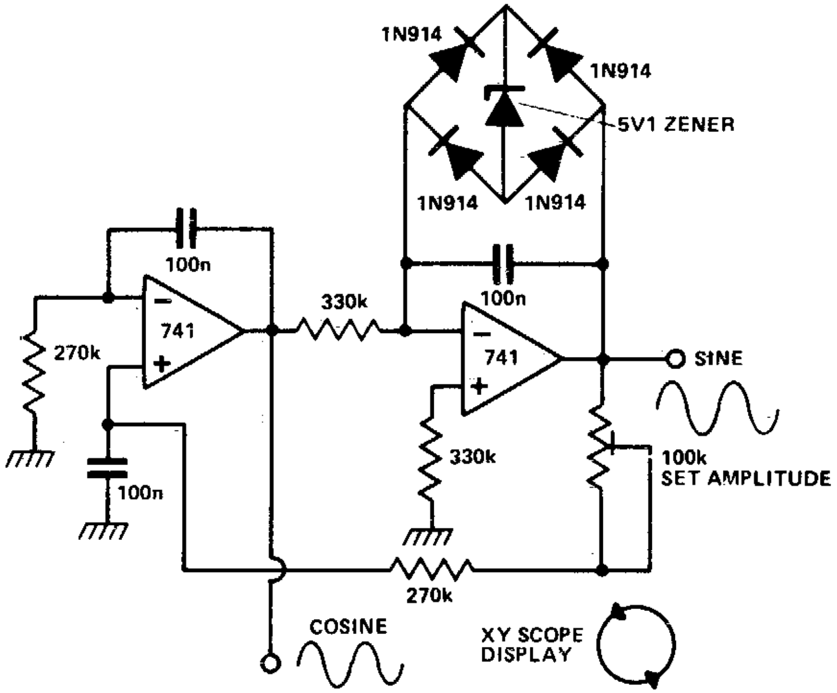

1N914
1N914
5V1 ZENER
1N914
1N914
100n
100n
330k
741
741
270k
SINE
330k
100k
SET AMPLITUDE
100n
270k
COSINE
XY SCOPE
DISPLAY

Fig. 60-4

Circuit Notes

The circuit is that of a quadrature sine and cosine oscillator. To generate circular displays, connect the two outputs to the X and Y inputs.

TRANSMITTER-OSCILLOSCOPE COUPLER FOR CB SIGNALS

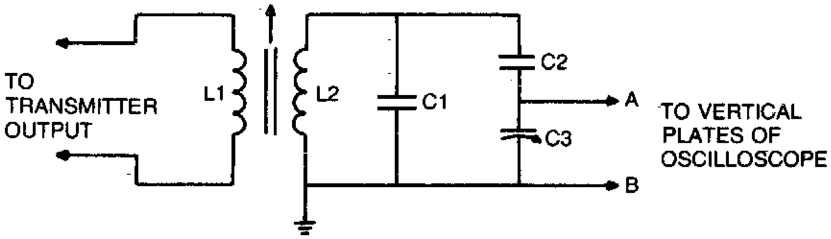
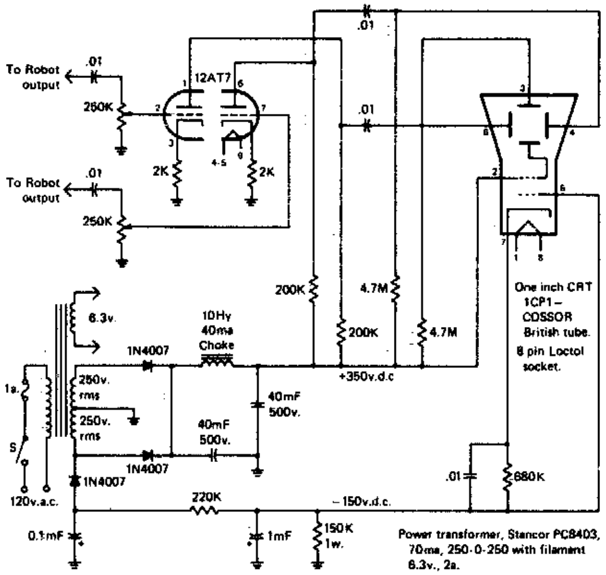

TO
TRANSMITTER
OUTPUT
L1
L2
C1
C2
C3
A
B
TO VERTICAL
PLATES OF
OSCILLOSCOPE

Fig. 60-5

Circuit Notes

To display an rf signal, connect L1 to the transmitter and points A and B to the vertical plates of the oscilloscope. Adjust L1 for minimum SWR and C3 for the desired trace height on the CRT. L2 = 4 turns #18 on ¾″ slug tuned rf coil form, L1 = 3 turns #22 adjacent to grounded end of L1, C1, and C2 = 5 pF, C3 = 75 pF trimmer.

OSCILLOSCOPE MONITOR

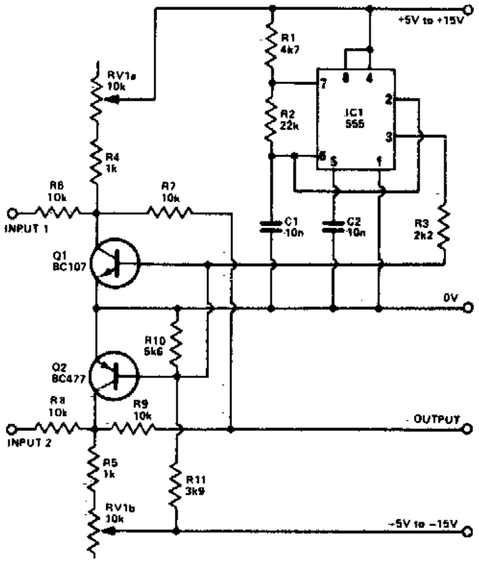


To Robot output
.01
250K
12AT7
2K
2K
To Robot output
.01
250K
.01
.01
200K
200K
4.7M
4.7M
One inch CRT 1CP1 – COSSOR British tube.
8 pin Loctol socket.
6.3v.
10Hy 40ma Choke
1N4007
250v. rms
250v. rms
1a.
S
120v.a.c.
40mF 500v.
40mF 500v.
+350v.d.c
1N4007
1N4007
.01
680K
220K
–150v.d.c.
0.1mF
1mF
150K 1w.

Power transformer, Stancor PC8403, 70ma, 250-0-250 with filament 6.3v., 2a.

Fig. 60-6

BEAM SPLITTER FOR OSCILLOSCOPE



+5V to +15V
R1 4k7
R2 22k
IC1 555
RV1a 10k
R4 1k
R6 10k
R7 10k
INPUT 1
Q1 BC107
C1 10n
C2 10n
R3 2k2
0V
R10 5k6
Q2 BC477
R8 10k
R9 10k
INPUT 2
OUTPUT
R5 1k
R11 3k9
RV1b 10k
–5V to –15V

Fig. 60-7

Circuit Notes

The basis of the beam-splitter is a 555 timer connected as an astable multivibrator. Signals at the two inputs are alternately displayed on the oscilloscope with a clear separation between them. The output is controlled by the tandem potentiometer RV1a/b which also varies the amplitude of the traces.

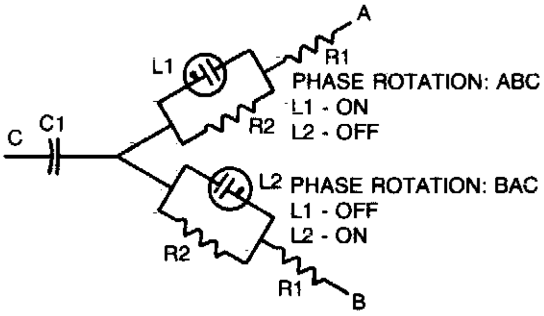
61

Phase Sequence and Phase Shift Circuits

The sources of the following circuits are contained in the Sources section beginning on page 730. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

Phase Sequence Indicator
Single Transistor Phase Shifter
0° to 180° Phase Shifter
Phase Shift Circuits
Precision Phase Splitter
0 to 360° Phase Shifter

PHASE SEQUENCE INDICATOR



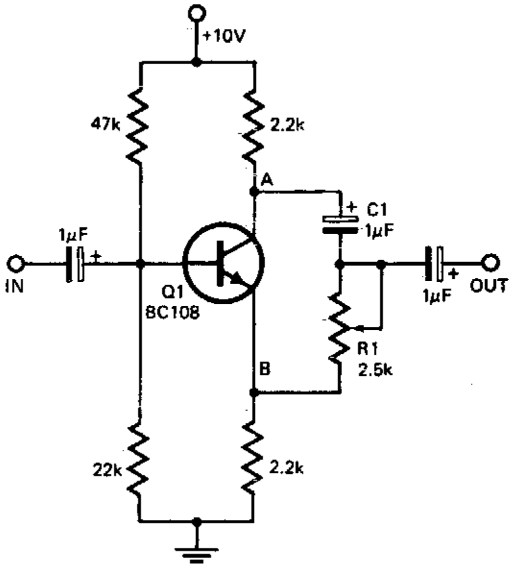
A
R1
L1
PHASE ROTATION: ABC
L1 - ON
R2
L2 - OFF
C
C1
L2
PHASE ROTATION: BAC
L1 - OFF
L2 - ON
R2
R1
B

Circuit Notes

Simple, portable phase-sequence indicator determines the proper phase rotation in polyphase circuits. Major components are two neon lamps, two resistors, and a capacitor. In operation, the leg voltages are unbalanced, so that the lamp with the maximum voltage—or proper phase sequence—lights. Table shows typical component values for various circuit frequencies.

Fig. 61-1

SINGLE TRANSISTOR PHASE SHIFTER



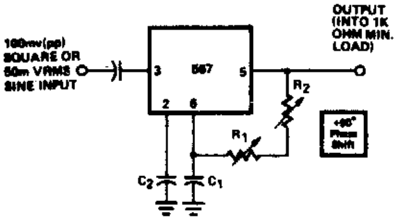
+10V
47k
2.2k
A
+ C1
1µF
1µF
+
IN
Q1
BC108
+
1µF
OUT
R1
2.5k
B
22k
2.2k

Fig. 61-2

Circuit Notes

This circuit provides a simple means of obtaining phase shifts between zero and 170°. The transistor operates as a phase splitter, the output at point A being 180° out of phase with the input. Point B is in phase with the input phase. Adjusting R1 provides the sum of various proportions of these and hence a continuously variable phase shift is provided. The circuit operates well in the 600 Hz to 4 kHz range.

0° TO 180° PHASE SHIFTER

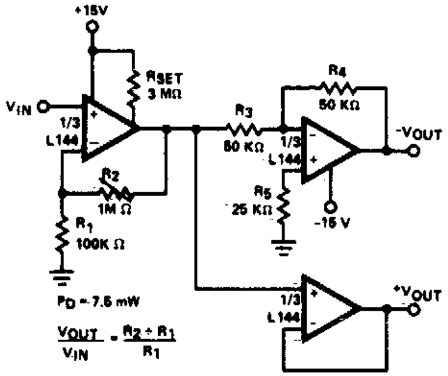

100mv(pp)
SQUARE OR
50m VRMS
SINE INPUT
3
567
5
2
6
OUTPUT
(INTO 1K
OHM MIN.
LOAD)
R2
R1
+90°
Phase
Shift
C2
C1

$R_2 = R_1/5$

Adjust R_1 so that $\phi = 90°$ with control midway

Fig. 61-3

PRECISION PHASE SPLITTER

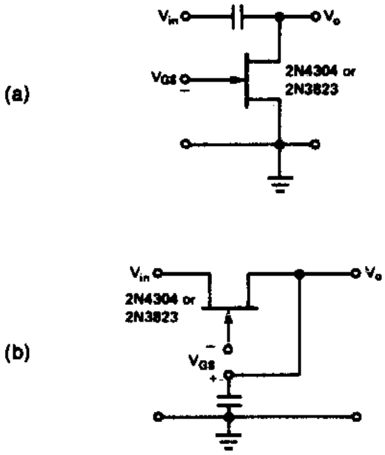

+15V
RSET
3 MΩ
VIN
1/3
L144
R3
50 KΩ
R4
50 KΩ
1/3
L144
-VOUT
R2
1M Ω
R1
100K Ω
R5
25 KΩ
-15 V
1/3
L144
+VOUT

$P_D = 7.5$ mW

$$\frac{V_{OUT}}{V_{IN}} = \frac{R_2 + R_1}{R_1}$$

Fig. 61-5

PHASE SHIFT CIRCUITS


(a)
Vin
Vo
VGS
2N4304 or
2N3823
(b)
Vin
Vo
2N4304 or
2N3823
VGS

(a) Phase advance circuit.

(b) Phase retard circuit.

Fig. 61-4

0° TO 360° PHASE SHIFTER

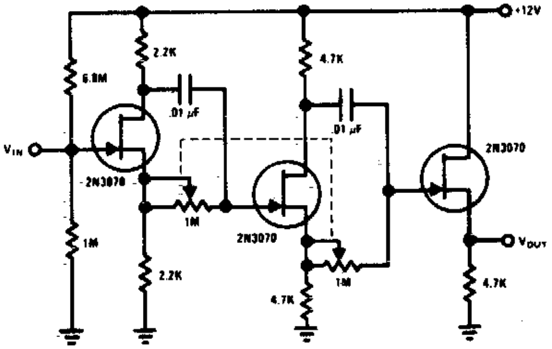

+12V
2.2K
6.8M
.01 μF
4.7K
.01 μF
VIN
2N3070
1M
1M
2N3070
2.2K
2N3070
4.7K
1M
VOUT
4.7K

Fig. 61-6

Circuit Notes

Each stage provides 0° to 180° phase shift. By ganging the two stages, 0° to 360° phase shift is achieved. The 2N3070 JFETs do not load the phase shift networks.

62

Photography Related Circuits

The sources of the following circuits are contained in the Sources section beginning on page 730. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

- Automatic Contrast Meter
- Darkroom Timer
- Photo Stop Action
- Sound Light-Flash Trigger
- Sound Activated Strobe Trip
- Flash Slave Driver
- Remote Flash Trigger
- Flash Exposure Meter
- Shutter Tester
- Photographic Timer

AUTOMATIC CONTRAST METER

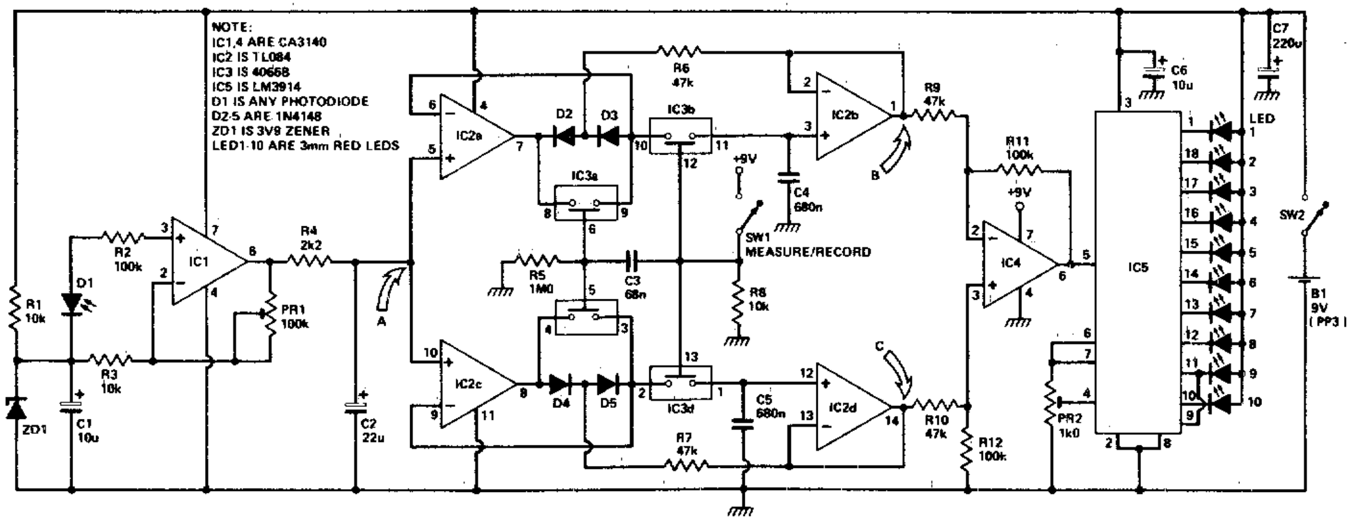

NOTE:
IC1,4 ARE CA3140
IC2 IS TL084
IC3 IS 4066B
IC5 IS LM3914
D1 IS ANY PHOTODIODE
D2-5 ARE 1N4148
ZD1 IS 3V9 ZENER
LED1-10 ARE 3mm RED LEDS
R1 10k
R2 100k
R3 10k
R4 2k2
PR1 100k
ZD1
D1
C1 10u
C2 22u
IC1
A
B
C
IC2a
IC2b
IC2c
IC2d
D2
D3
D4
D5
IC3a
IC3b
IC3d
R5 1M0
C3 68n
R6 47k
R7 47k
C4 680n
C5 680n
+9V
SW1 MEASURE/RECORD
R8 10k
R9 47k
R10 47k
R11 100k
R12 100k
IC4
IC5
PR2 1k0
C6 10u
C7 220u
LED
SW2
B1 9V (PP3)

Fig. 62-1

Circuit Notes

The circuit arrangement consists of a photo-amplifier which feeds a voltage derived from varying light levels in an enlarger to a pair of peak detectors. One follows the peak positive voltage and the other the peak negative voltage. The capacitors used for storing the voltage peaks in the followers also form part of sample and hold circuits which are then switched to hold after the measurement. Their outputs represent the maximum and minimum values of light intensity. A differential amplifier then computes the ratio of these values, and the result is displayed on an LED bargraph meter.

DARKROOM TIMER

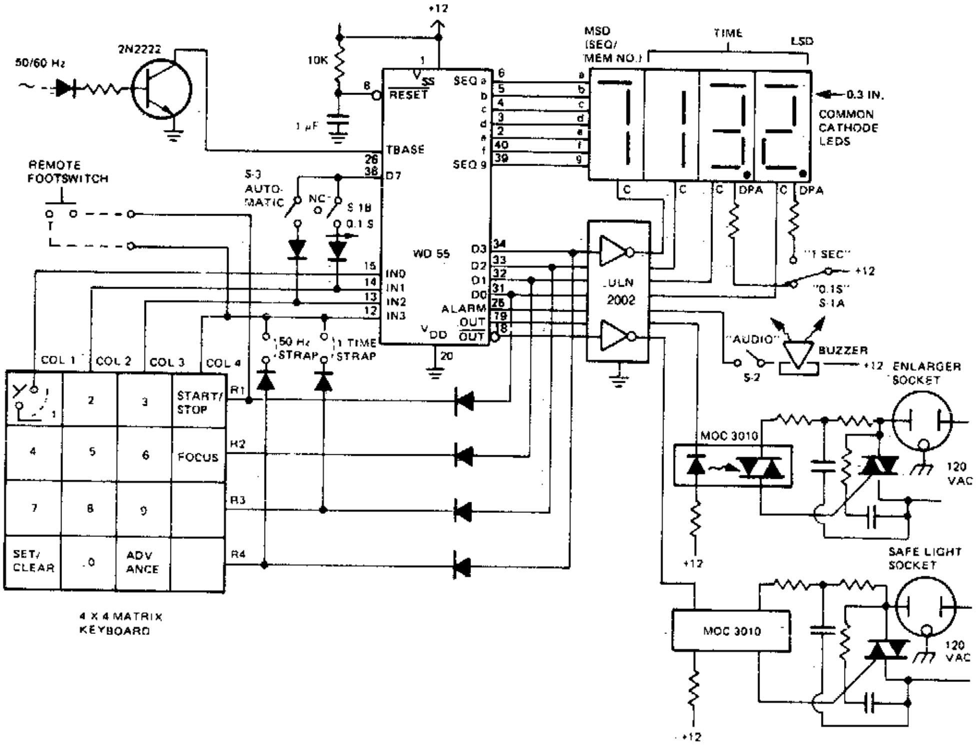

+12
50/60 Hz
2N2222
10K
1 µF
8
RESET
1
V_{SS}
SEQ a
b
c
d
e
f
SEQ 9
TBASE
26
38
D7
REMOTE FOOTSWITCH
S-3 AUTO-MATIC
NC
S 1B
0.1 S
WD 55
15
14
13
12
IN0
IN1
IN2
IN3
D3
D2
D1
D0
ALARM
OUT
V_{DD}
20
50 Hz STRAP
1 TIME STRAP
COL 1
COL 2
COL 3
COL 4
START/ STOP
FOCUS
SET/ CLEAR
.0
ADV ANCE
R1
R2
R3
R4
4 X 4 MATRIX KEYBOARD
MSD (SEQ/ MEM NO.)
TIME
LSD
0.3 IN.
COMMON CATHODE LEDS
C
DPA
"1 SEC"
"0.1S"
S-1A
ULN 2002
"AUDIO"
S-2
BUZZER
+12 ENLARGER SOCKET
MOC 3010
120 VAC
SAFE LIGHT SOCKET
+12

Fig. 62-2

Circuit Notes

The darkroom timer/controller uses few external components: a display, a digit driver, keyboard, and output switching devices. A 4-digit common-cathode LED display is desirable for dark room environments. The time base is provided by shaping up the 50/60 Hz ac line. A DPDT switch (S1) is used to select a resolution of .1 or 1 seconds and to simultaneously move the decimal point. Timer/controller has two switched ac outlets, one for the enlarger and one for the safe light. They are the complements of each other in that the safe light is on when the enlarger is not active and is off when the enlarger is printing. The buzzer is of the self-contained oscillator variety and operates with dc drive.

PHOTO STOP ACTION

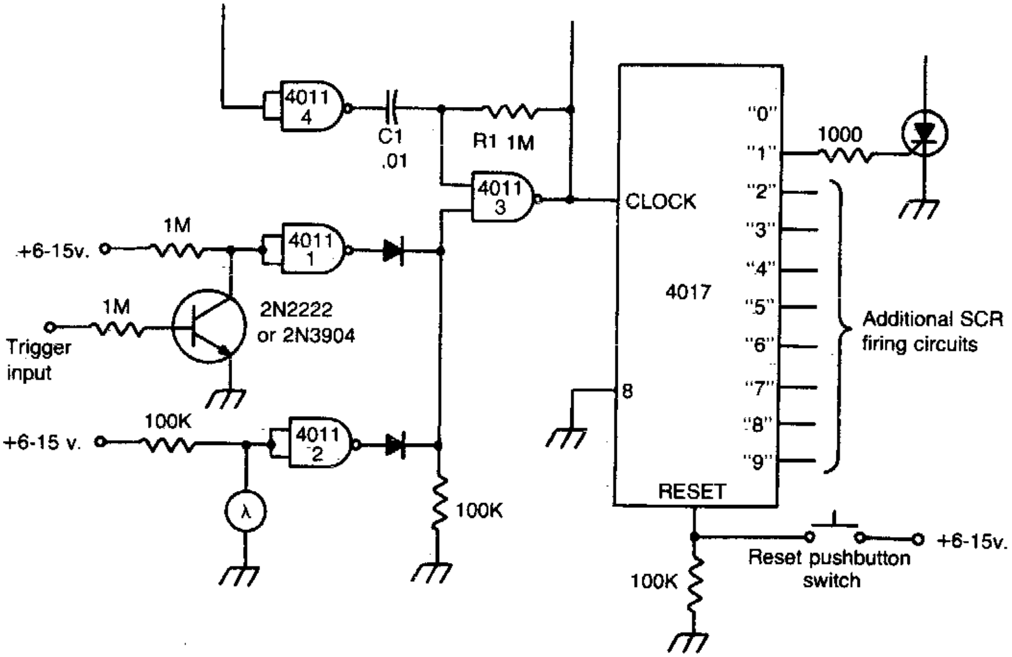

4011
4
C1
.01
R1 1M
4011
3
CLOCK
4017
"0"
"1"
"2"
"3"
"4"
"5"
"6"
"7"
"8"
"9"
1000
Additional SCR firing circuits
1M
+6-15v.
4011
1
1M
Trigger input
2N2222
or 2N3904
100K
+6-15 v.
4011
2
100K
8
RESET
+6-15v.
Reset pushbutton switch
100K

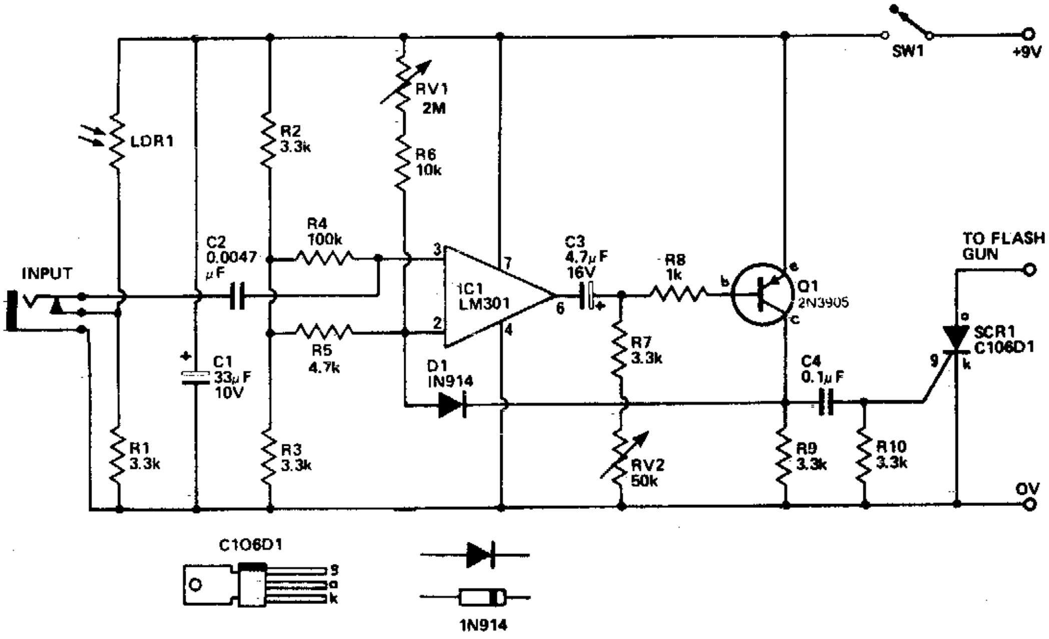
Fig. 62-3

Bulb firing system SCR

Circuit Notes

This circuit gives multiple "stop-action" photographic effects like showing a bouncing ball in up to nine locations in a single photograph. The circuit will automatically fire the bulbs sequentially with the time between each firing variable. The circuit is functionally complete except for the actual firing system. In many cases, a simple SCR will work, as shown. The firing can be initiated in one of two ways. A trigger pulse can be applied to the trigger input terminal through a capacitor, or can operate the unit as a slave. Light from a camera-mounted flash will activate the circuit through its built-in photocell pickup. The time period between each successive flash is determined by C1 and R1, which is variable. After firing the circuit, it must be reset by momentarily depressing the reset button.

SOUND LIGHT-FLASH TRIGGER



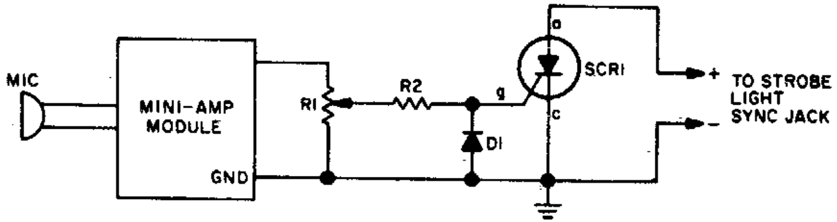
SW1
+9V
LDR1
R2
3.3k
RV1
2M
R6
10k
R4
100k
C2
0.0047
μF
INPUT
3
2
IC1
LM301
7
6
4
C3
4.7μF
16V
R8
1k
b
e
c
Q1
2N3905
TO FLASH
GUN
SCR1
C106D1
a
g
k
C1
33μF
10V
R5
4.7k
D1
IN914
R7
3.3k
C4
0.1μF
R1
3.3k
R3
3.3k
RV2
50k
R9
3.3k
R10
3.3k
0V
C106D1
g
a
k
1N914

Fig. 62-4

Circuit Notes

Sound input to the microphone triggers the IC monostable circuit which subsequently triggers an SCR, and hence the flash, after a time delay. This delay is adjustable—by varying the monostable on-time—from from 5 milliseconds to 200 milliseconds.

SOUND ACTIVATED STROBE TRIP


MIC
MINI-AMP MODULE
GND
R1
R2
g
SCR1
a
c
D1
+ TO STROBE LIGHT SYNC JACK
-

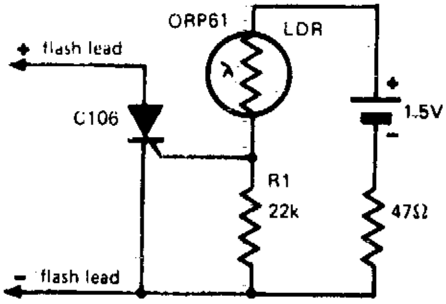
D1—HEP-154 silicon rectifier
R1—5000-ohm potentiometer
R2—2700-ohm, ½-watt resistor
SCR1— silicon- controlled rectifier
MIC.—Ceramic microphone

Fig. 62-5

Circuit Notes

Take strobe-flash pictures the instant a pin pricks a balloon, a hammer breaks a lamp bulb or a bullet leaves a gun. Use a transistor amplifier of 1-watt rating or less. (It must have an output transformer.) The amplifier is terminated with a resistor on its highest output impedance, preferably 16 ohms. To test, darken room lights, open camera shutter, and break a lamp bulb with a hammer. The sound of the hammer striking the lamp will trigger the flash, and the picture will have been taken at that instant.

FLASH SLAVE DRIVER


ORP61
LDR
+ flash lead
C106
+
1.5V
-
R1
22k
47Ω
- flash lead

Circuit Notes

In photography, a separate flash, triggered by the light of a master flash light, is often required to provide more light, fill-in shadows etc. The sensitivity of this circuit depends on the proximity of the master flash and the value of R1. Increasing R1 gives increased sensitivity.

Fig. 62-6

REMOTE FLASH TRIGGER

Q1—300-V light-activated silicon-controlled rectifier (LASCR)
R1—47,000-ohm, ½-watt resistor

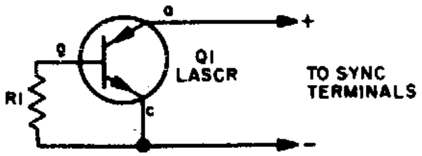

a
+
g
Q1
LASCR
TO SYNC
TERMINALS
R1
c
−

Fig. 62-7

Circuit Notes

Transistor Q1 is a light-activated silicon-controlled rectifier (LASCR). The gate is tripped by light entering a small lens built into the top cap. To operate, provide a 6-in. length of stiff wire for the anode and cathode connections and terminate the wires in a polarized power plug that matches the sync terminals on your electronic flashgun (strobelight). Make certain the anode lead connects to the positive sync terminal. When using the device, bend the connecting wires so the LASCR lens faces the main flash. This will fire the remote unit.

FLASH EXPOSURE METER

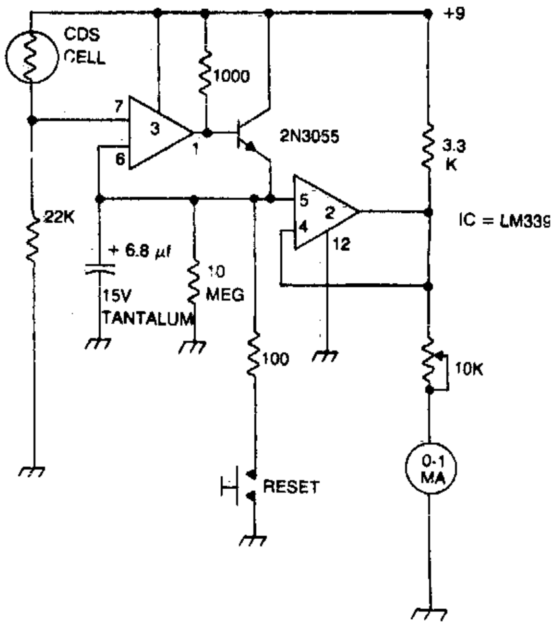

+9
CDS
CELL
1000
7
3
6
1
2N3055
3.3
K
5
2
4
12
IC = LM339
22K
+ 6.8 μf
15V
TANTALUM
10
MEG
100
10K
0-1
MA
RESET

Fig. 62-8

Circuit Notes

Strobe light meter catches the peak of flash intensity and holds it long enough to give a reading. The reset button must be pressed before each measurement.

SHUTTER TESTER

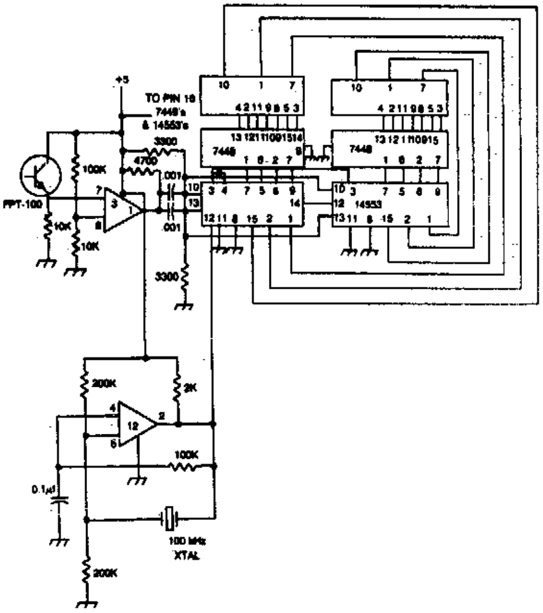

+5
TO PIN 16
7448's
& 14553's
3300
4700
.001
.001
100K
10K
10K
FPT-100
3300
7448
7448
14553
200K
2K
12
100K
0.1µf
100 MHz
XTAL
200K

Fig. 62-9

Circuit Notes

Shutter speed tester combines frequency counter, crystal oscillator, and phototransistor-operated gate generator. Oscillator pulses are counted as long as the shutter is open. Reset is automatic at the instant the shutter opens.

PHOTOGRAPHIC TIMER

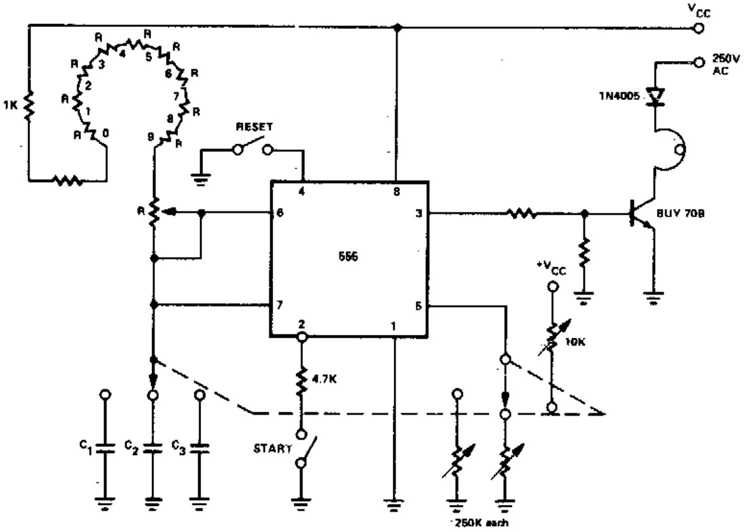

V_{CC}
250V
AC
1N4005
1K
R
RESET
556
BUY 708
$+V_{CC}$
10K
4.7K
START
C_1
C_2
C_3
250K each

Fig. 62-10

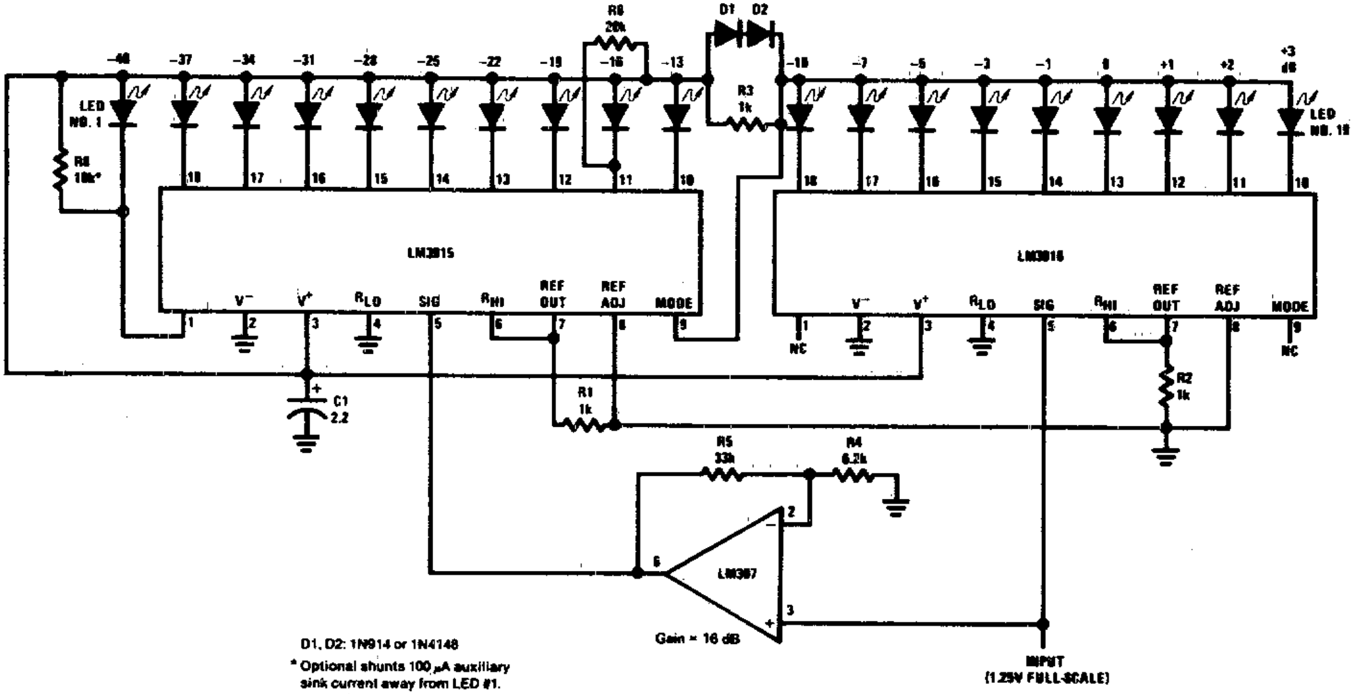
63

Power Measuring Circuits

The sources of the following circuits are contained in the Sources section beginning on page 730. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

Extended Range VU Meter (Dot Mode)
Audio Power Meter
Audio Power Meter
Power Meter (1 kW Full Scale)
60 MHz Power Gain Test Circuit

EXTENDED RANGE VU METER (DOT MODE)


R6
20k
D1 D2
−40
−37
−34
−31
−28
−25
−22
−19
−16
−13
−10
−7
−5
−3
−1
0
+1
+2
+3
dB
LED
NO. 1
R8
10k*
R3
1k
LED
NO. 19
18
17
16
15
14
13
12
11
10
LM3915
LM3916
V⁻
V⁺
R_LO
SIG
R_HI
REF
OUT
REF
ADJ
MODE
1
2
3
4
5
6
7
8
9
NC
NC
C1
2.2
R1
1k
R2
1k
R5
33k
R4
6.2k
LM387
6
2
3
Gain = 16 dB
INPUT
(1.25V FULL-SCALE)

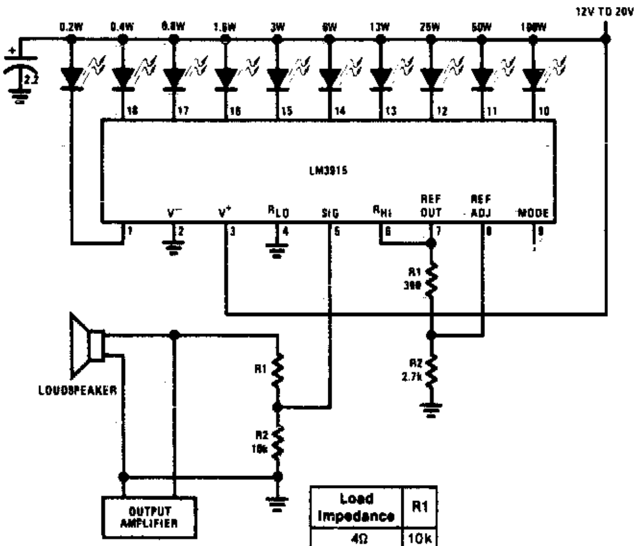
D1, D2: 1N914 or 1N4148

* Optional shunts 100 μA auxiliary sink current away from LED #1.

†See Application Hints for optional peak or average detector

Fig. 63-1

AUDIO POWER METER


12V TO 20V
0.2W
0.4W
0.8W
1.6W
3W
6W
13W
25W
50W
100W
2.2
18
17
16
15
14
13
12
11
10
LM3915
V⁻
V⁺
RLO
SIG
RHI
REF OUT
REF ADJ
MODE
1
2
3
4
5
6
7
8
9
R1 390
R2 2.7k
LOUDSPEAKER
R1
R2 10k
OUTPUT AMPLIFIER

Load Impedance	R1
4Ω	10k
8Ω	18k
16Ω	30k

See Application Hints for optional Peak or Average Detector

Fig. 63-2

AUDIO POWER METER

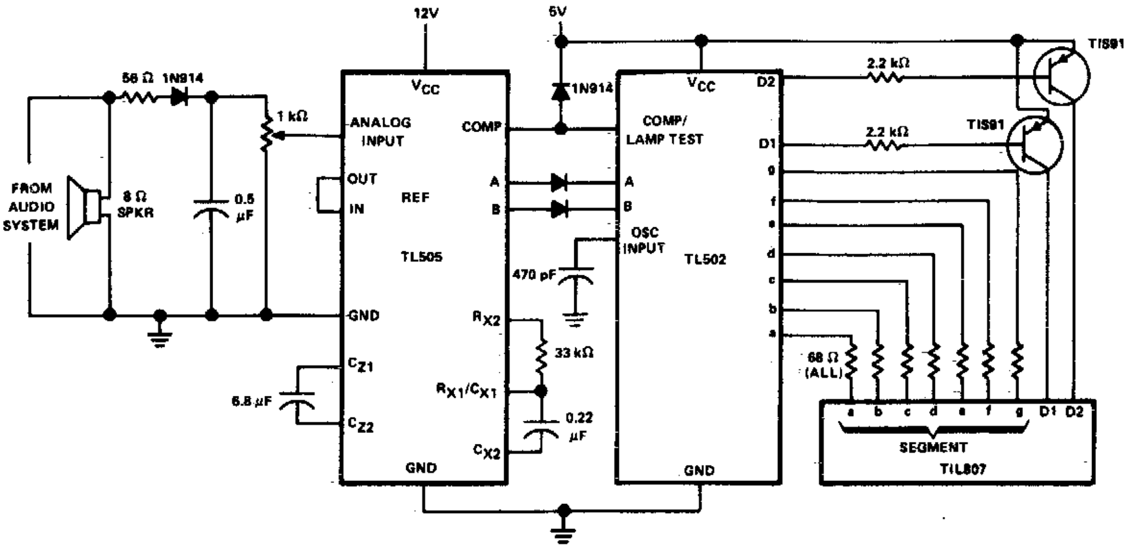

12V
5V
56 Ω
1N914
1 kΩ
FROM AUDIO SYSTEM
8 Ω SPKR
0.5 µF
VCC
ANALOG INPUT
COMP
OUT
IN
REF
A
B
TL505
GND
RX2
CZ1
CZ2
RX1/CX1
CX2
GND
6.8 µF
33 kΩ
0.22 µF
470 pF
1N914
VCC
COMP/ LAMP TEST
OSC INPUT
TL502
GND
D2
D1
g
f
e
d
c
b
a
2.2 kΩ
2.2 kΩ
TIS91
TIS91
68 Ω (ALL)
a b c d e f g D1 D2
SEGMENT
TIL807

Fig. 63-3

POWER METER (1 kW FULL SCALE)

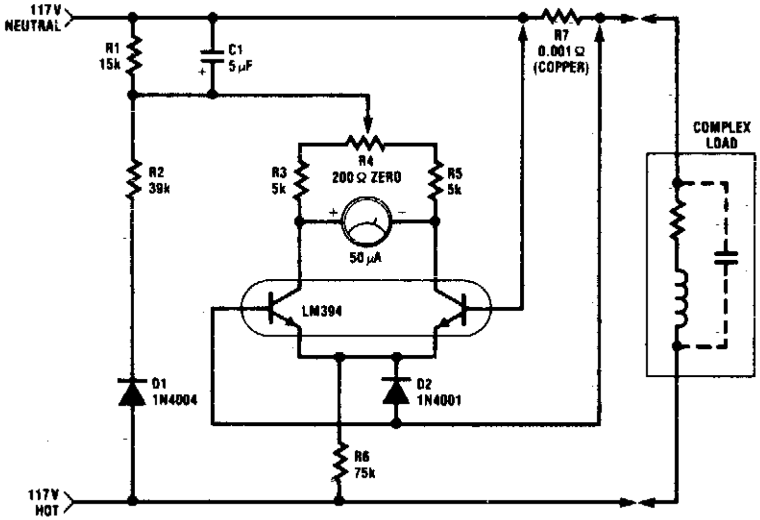
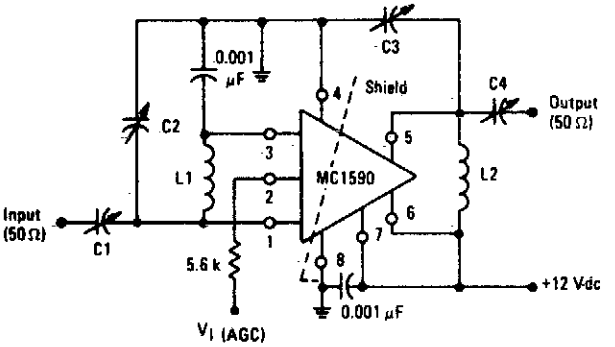

117V NEUTRAL
R1 15k
C1 5 µF
+
R7 0.001 Ω (COPPER)
COMPLEX LOAD
R4 200 Ω ZERO
R3 5k
R5 5k
R2 39k
+
−
50 µA
LM394
D1 1N4004
D2 1N4001
R6 75k
117V HOT

Fig. 63-4

Circuit Notes

The circuit is intended for 117 Vac ± 50 Vac operation, but can be easily modified for higher or lower voltages. It measures true (nonreactive) power being delivered to the load and requires no external power supply. Idling power drain is only 0.5 W. Load current sensing voltage is only 10 mV, keeping load voltage loss to 0.01%. Rejection of reactive load currents is better than 100:1 for linear loads. Nonlinearity is about 1% full scale when using a 50 µA meter movement.

60 MHz POWER GAIN TEST CIRCUIT


0.001 µF
C3
4
Shield
C4
Output (50 Ω)
C2
3
5
L1
MC1590
L2
2
Input (50 Ω)
C1
1
6
7
5.6 k
8
+12 Vdc
0.001 µF
V_I (AGC)

L1 = 7 Turns, #20 AWG Wire, 5/16" Dia., 5/8" Long
L2 = 6 Turns, #14 AWG Wire, 9/16" Dia., 3/4" Long
C1,C2,C3 = (1-30) pF
C4 = (1-10) pF

Fig. 63-5

64

Power Supplies (Fixed)

The sources of the following circuits are contained in the Sources section beginning on page 730. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

Switching Regulator Operating at 200 kHz
5 V, 0.5 A Power Supply
3 W Switching Regulator Application Circuit
Regulated Split Supplies from a Single Supply
Switching Step-Down Regulator
Single-Ended Regulator
±50 V Push-Pull Switched Mode Converter
5 V/0.5 A Buck Converter
±50 V Feed Forward Switch Mode Converter
Traveller's Shaver Adapter
100 Vrms Voltage Regulator
Transistor Increases Zener Rating
Dual Polarity Power Supply
5.0 V/6.0 A, 25 kHz Switching Regulator with Separate Ultra-Stable Reference
Mobile Voltage Regulator
Negative Switching Regulator
Positive Switching Regulator
Positive Floating Regulator
Negative Floating Regulator
Negative Voltage Regulator
−15 V Negative Regulator
Slow Turn-On 15 V Regulator
High Stability 10 V Regulator
5 V/1 A Switching Regulator
15 V/1 A Regulator with Remote Sense
Low Ripple Power Supply
5.0 V/10 A Regulator
5.0 V/3.0 A Regulator
100 V/10.25 A Switch Mode Converter
Voltage Regulator
Low Voltage Regulators with Short Circuit Protection
High Stability 1 A Regulator
100 V/0.25 A Switch Mode Converter

SWITCHING REGULATOR OPERATING AT 200 kHz

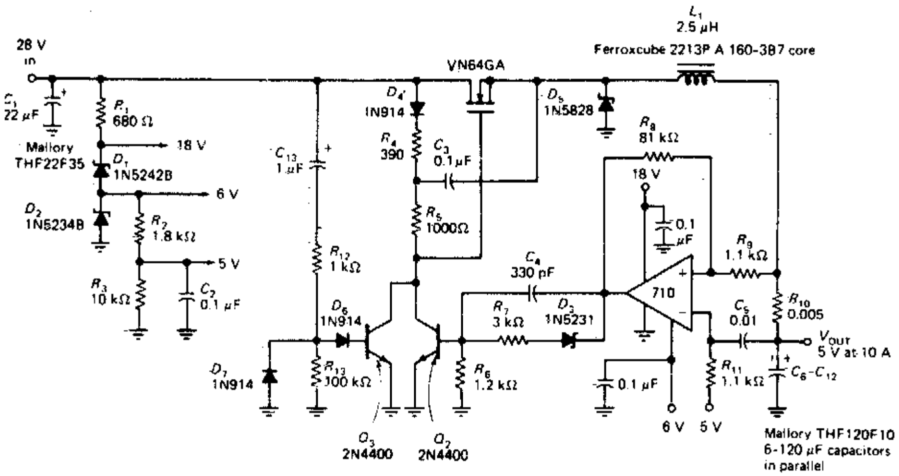

28 V in
C_1 22 μF
Mallory THF22F35
R_1 680 Ω
18 V
D_1 1N5242B
6 V
D_2 1N5234B
R_2 1.8 kΩ
5 V
R_3 10 kΩ
C_2 0.1 μF
C_{13} 1 μF
R_{12} 1 kΩ
D_6 1N914
D_7 1N914
R_{13} 100 kΩ
D_4 1N914
R_4 390
C_3 0.1 μF
R_5 1000Ω
VN64GA
Q_3 2N4400
Q_2 2N4400
R_6 1.2 kΩ
C_4 330 pF
R_7 3 kΩ
D_3 1N5231
D_5 1N5828
L_1 2.5 μH
Ferroxcube 2213P A 160-387 core
R_8 81 kΩ
18 V
0.1 μF
710
0.1 μF
6 V
5 V
R_9 1.1 kΩ
R_{10} 0.005
C_5 0.01
R_{11} 1.1 kΩ
C_6-C_{12}
V_{OUT} 5 V at 10 A
Mallory THF120F10 6-120 μF capacitors in parallel

Fig. 64-1

Circuit Notes

This circuit provides a regulated dc with less than 100 mV of ripple for microprocessor applications. Necessary operating voltages are taken from the bleeder resistor network connected across the unregulated 28 V supply. The output of the LM710 comparator (actually an oscillator running at 200 kHz) is fed through a level-shifting circuit to the base of bipolar transistor Q2. This transistor is part of a bootstrap circuit necessary to turn the power MOSFET full on in totem-pole MOSFET arrays.

5 V, 0.5 A POWER SUPPLY

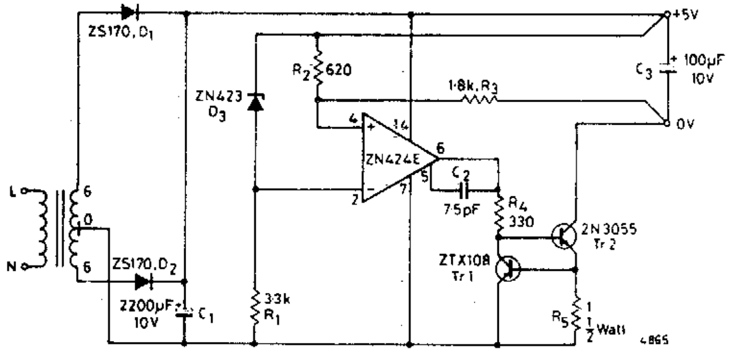
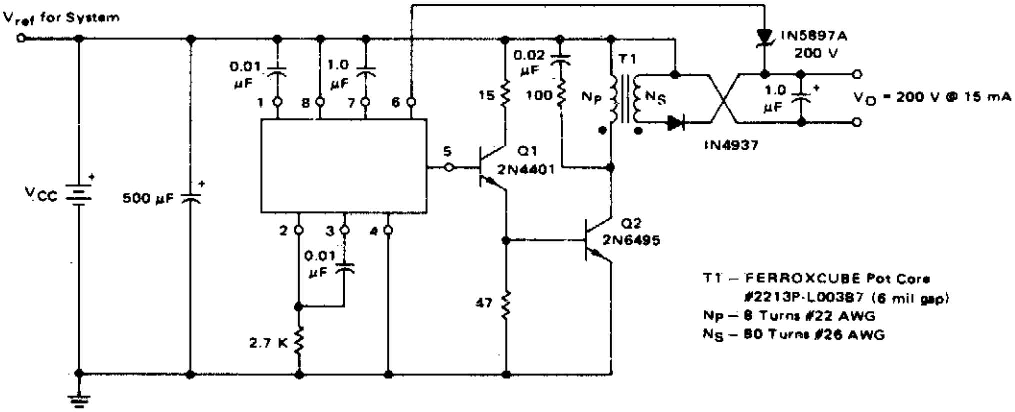

ZS170, D_1
+5V
R_2 620
C_3 100μF 10V
ZN423 D_3
1·8k, R_3
0V
4
14
6
ZN424E
2
7
5
C_2
7·5pF
R_4 330
2N3055 Tr 2
ZTX108 Tr 1
L
N
6
0
6
ZS170, D_2
2200μF 10V
C_1
3·3k R_1
R_5 1 ½ Watt
4865

Fig. 64-2

Circuit Notes

The circuit is essentially a constant source modified by the feedback components R2 and R3 to give a constant voltage output. The output of the ZN424E need only be 2 volts above the negative rail, by placing the load in the collector of the output transistor Tr2. The current circuit is achieved by Tr1 and R5. This simple circuit has the following performance characteristics: Output noise and ripple (full load) = 1 mV rms. Load regulation (0 to 0.5 A) = 0.1%. Temperature coefficient = ± 100 ppm/°C. Current limit = 0.65 A.

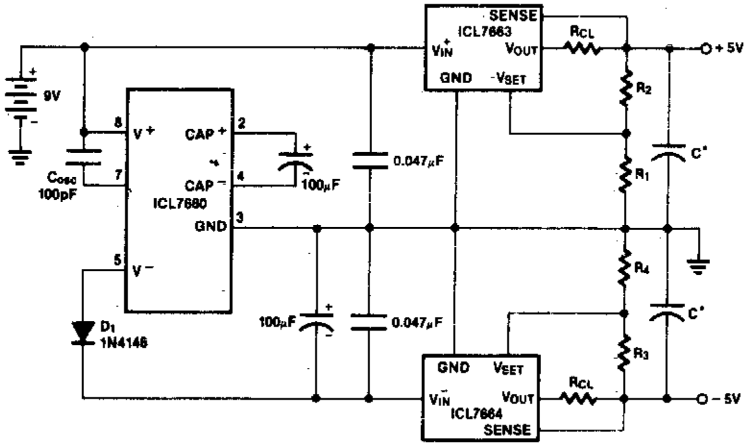
3 W SWITCHING REGULATOR APPLICATION CIRCUIT


V_ref for System
VCC
500 µF
0.01 µF
1.0 µF
1
8
7
6
5
2
3
4
0.01 µF
2.7 K
15
100
0.02 µF
47
Q1 2N4401
Q2 2N6495
T1
Np
Ns
1N5897A 200 V
1N4937
1.0 µF
V_O = 200 V @ 15 mA
T1 – FERROXCUBE Pot Core #2213P-L00387 (6 mil gap)
Np – 8 Turns #22 AWG
Ns – 80 Turns #26 AWG

3-Watt Switching Regulator - converts 5 V to 200 V for gas discharge displays such as Burroughs Panaplex and Beckman.

Fig. 64-3

REGULATED SPLIT POWER SUPPLIES FROM A SINGLE SUPPLY


9V
Cosc 100pF
8
7
5
V+
V−
CAP+
CAP−
GND
ICL7660
2
4
3
100µF
0.047µF
D1 1N4148
100µF
0.047µF
SENSE
ICL7663
VIN
VOUT
GND
VSET
RCL
R2
R1
R4
R3
C*
C*
+5V
−5V
GND
VSET
VIN
VOUT
ICL7664
SENSE
RCL

*Values depend on load characteristics

Fig. 64-4

Circuit Notes

The oscillation frequency of the ICL7660 is reduced by the external oscillator capacitor, so that it inverts the battery voltage more efficiently.

SWITCHING STEP-DOWN REGULATOR

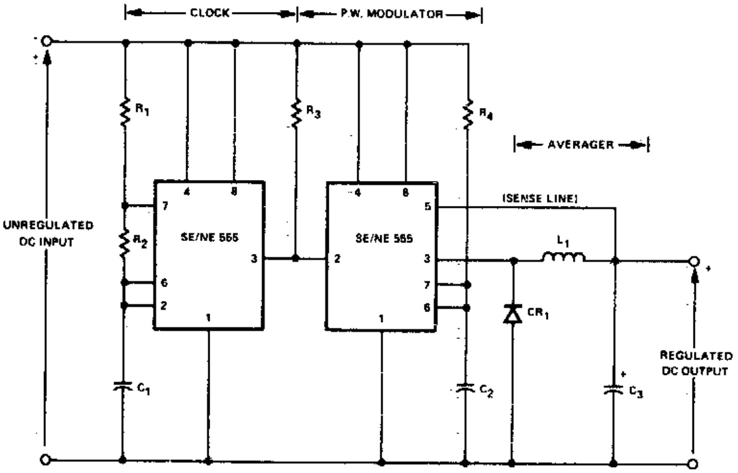

CLOCK
P.W. MODULATOR
AVERAGER
R1
R2
R3
R4
C1
C2
C3
SE/NE 555
SE/NE 555
4
8
7
6
2
3
1
5
(SENSE LINE)
L1
CR1
UNREGULATED DC INPUT
REGULATED DC OUTPUT
+

Fig. 64-5

SINGLE-ENDED REGULATOR

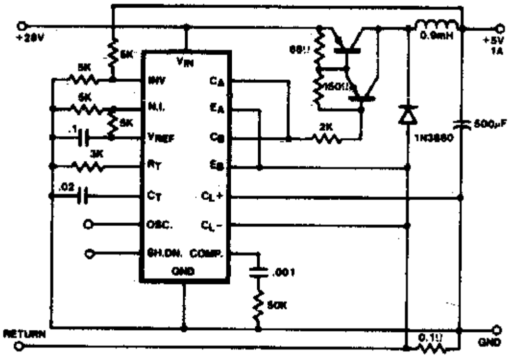
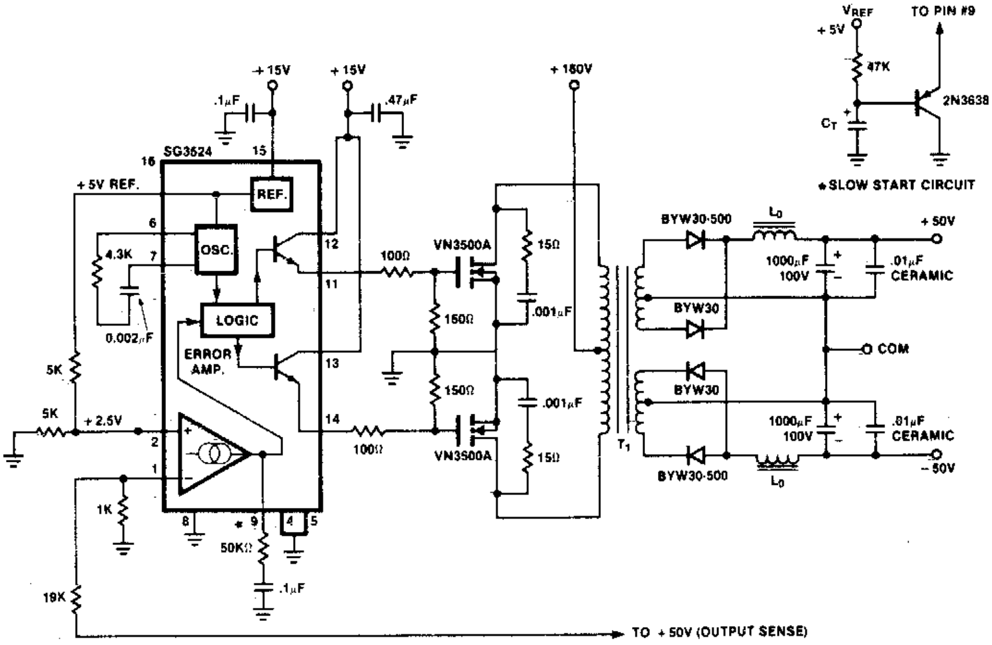

+28V
5K
5K
5K
5K
.1
3K
.02
VIN
INV
N.I.
VREF
RT
CT
OSC.
SH.DN.
COMP.
GND
CA
EA
CB
EB
CL+
CL−
68Ω
150Ω
2K
0.9mH
+5V
1A
1N3880
500µF
.001
50K
RETURN
0.1Ω
GND

Fig. 64-6

Circuit Notes

In this conventional single-ended regulator circuit, the two outputs of the SG1524 are connected in parallel for effective 0-90% duty-cycle modulation. The use of an output inductor requires an RC phase compensation network for loop stability.

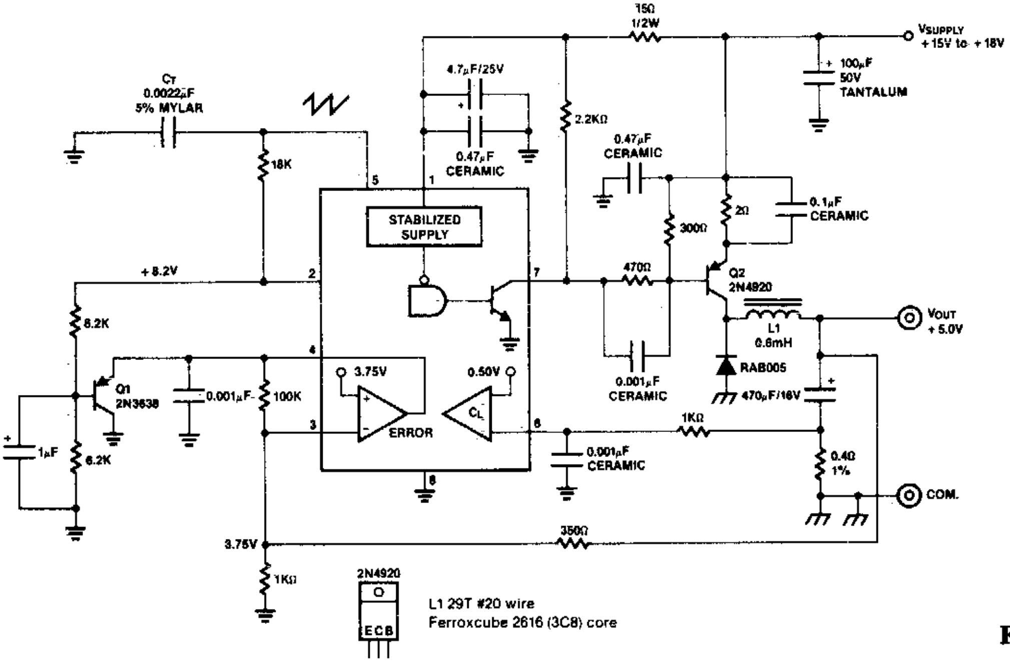
±50 V PUSH-PULL SWITCHED MODE CONVERTER



VREF
+5V
TO PIN #9
47K
CT
2N3638
* SLOW START CIRCUIT
+15V
+15V
+180V
.1μF
.47μF
SG3524
16
15
+5V REF.
REF.
6
OSC.
4.3K
7
0.002μF
12
11
LOGIC
ERROR AMP.
13
14
5K
5K
+2.5V
2
1
1K
8
* 9
4
5
50KΩ
.1μF
19K
TO +50V (OUTPUT SENSE)
100Ω
VN3500A
15Ω
.001μF
150Ω
150Ω
.001μF
100Ω
VN3500A
15Ω
T1
BYW30-500
L0
+50V
1000μF 100V
.01μF CERAMIC
BYW30
COM
BYW30
1000μF 100V
.01μF CERAMIC
BYW30-500
L0
−50V

Fig. 64-7

5 V/0.5 A BUCK CONVERTER



15Ω 1/2W
VSUPPLY +15V to +18V
100μF 50V TANTALUM
CT 0.0022μF 5% MYLAR
4.7μF/25V
0.47μF CERAMIC
2.2KΩ
18K
5
1
STABILIZED SUPPLY
0.47μF CERAMIC
2Ω
0.1μF CERAMIC
300Ω
+8.2V
2
7
470Ω
Q2 2N4920
8.2K
L1 0.6mH
VOUT +5.0V
4
3.75V
0.50V
Q1 2N3638
0.001μF
100K
RAB005
470μF/16V
1μF
6.2K
3
ERROR
CL
6
0.001μF CERAMIC
0.001μF CERAMIC
1KΩ
0.4Ω 1%
8
COM.
3.75V
350Ω
1KΩ
2N4920
ECB

L1 29T #20 wire
Ferroxcube 2616 (3C8) core

Fig. 64-8

±50 V FEED FORWARD SWITCH MODE CONVERTER

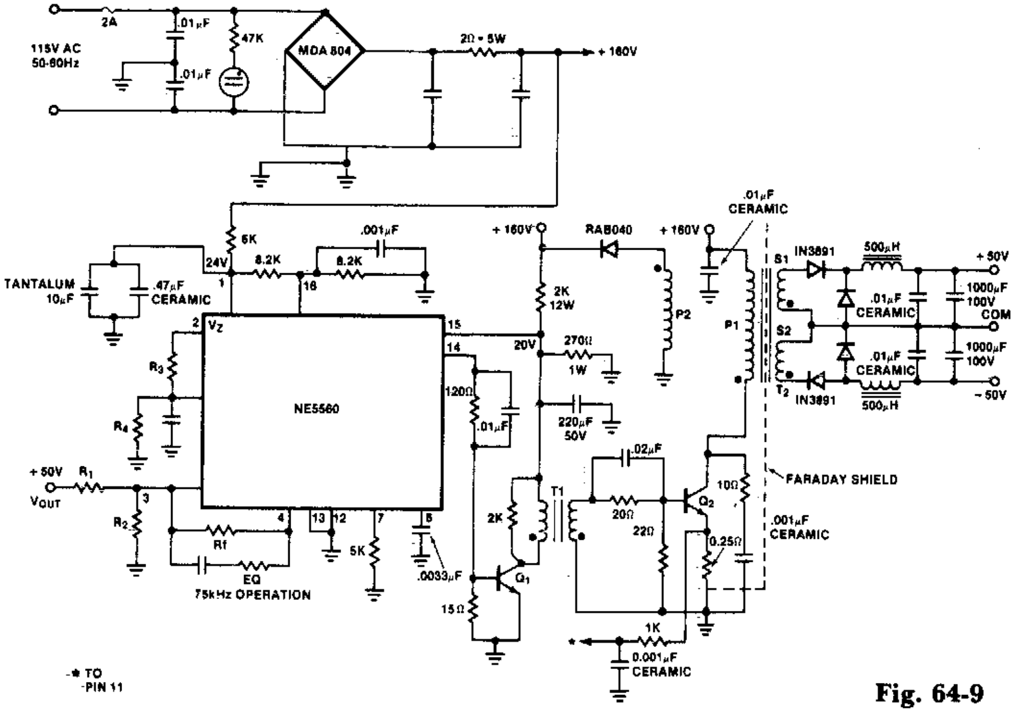

115V AC
50-60Hz
2A
.01μF
.01μF
47K
MDA 804
2Ω - 5W
+ 160V
TANTALUM
10μF
.47μF
CERAMIC
5K
24V
8.2K
8.2K
.001μF
NE5560
Vz
R3
R4
R1
R2
+ 50V
VOUT
Rf
EQ
75kHz OPERATION
5K
.0033μF
120Ω
.01μF
15Ω
Q1
2K
T1
+ 160V
RA8040
+ 160V
2K
12W
20V
270Ω
1W
220μF
50V
P2
P1
.02μF
20Ω
22Ω
Q2
10Ω
0.25Ω
1K
0.001μF
CERAMIC
.01μF
CERAMIC
.001μF
CERAMIC
FARADAY SHIELD
S1
S2
T2
IN3891
IN3891
500μH
500μH
.01μF
CERAMIC
.01μF
CERAMIC
1000μF
100V
1000μF
100V
+ 50V
COM
- 50V
* TO
PIN 11

Fig. 64-9

TRAVELLER'S SHAVER ADAPTER

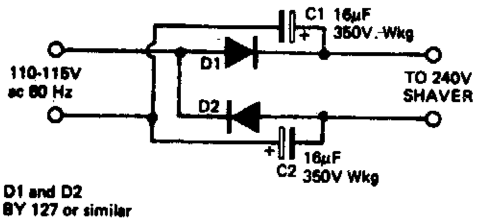
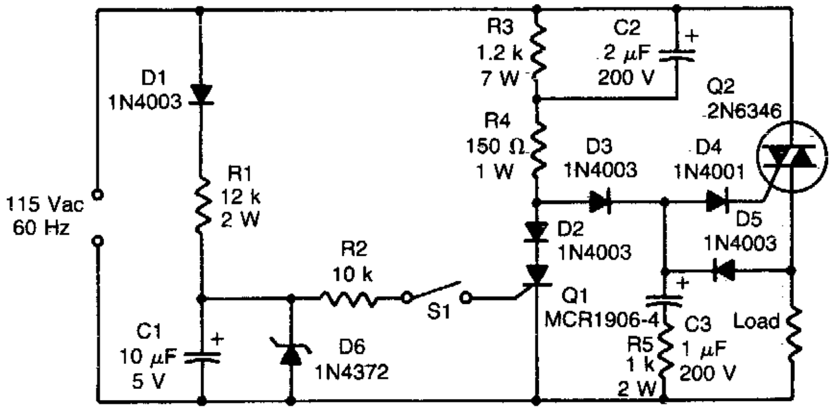

C1 16μF
350V. Wkg
D1
D2
110-115V
ac 60 Hz
TO 240V
SHAVER
16μF
C2 350V Wkg
D1 and D2
BY 127 or similar

Fig. 64-10

Circuit Notes

Many countries have 115 volts mains supplies. This can be a problem if your electric shaver is designed for 220/240 volts only. This simple rectifier voltage doubler enables motor driven 240 volt shavers to be operated at full speed from a 115 volt supply. As the output voltage is dc, the circuit can only be used to drive small ac/dc motors. It cannot be used, for example, to operate vibrator-type shavers, or radios unless the latter are ac/dc operated.

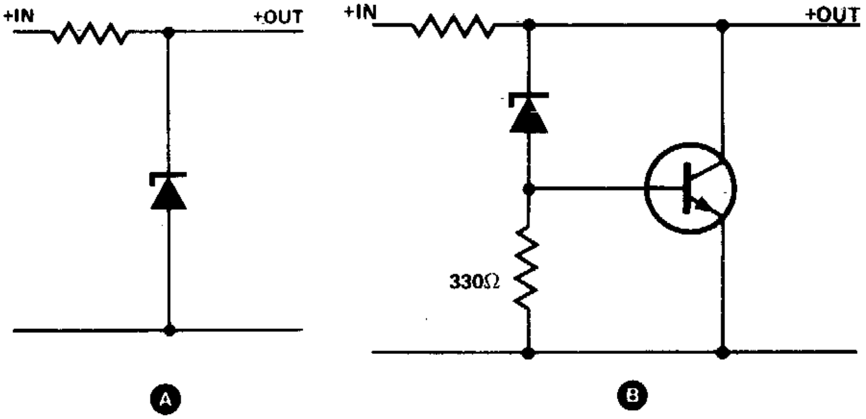
100 Vrms VOLTAGE REGULATOR



115 Vac
60 Hz
D1
1N4003
R1
12 k
2 W
C1
10 μF
5 V
D6
1N4372
R2
10 k
S1
R3
1.2 k
7 W
R4
150 Ω
1 W
C2
2 μF
200 V
Q2
2N6346
D3
1N4003
D4
1N4001
D2
1N4003
D5
1N4003
Q1
MCR1906-4
C3
1 μF
200 V
R5
1 k
2 W
Load

Fig. 64-11

TRANSISTOR INCREASES ZENER RATING



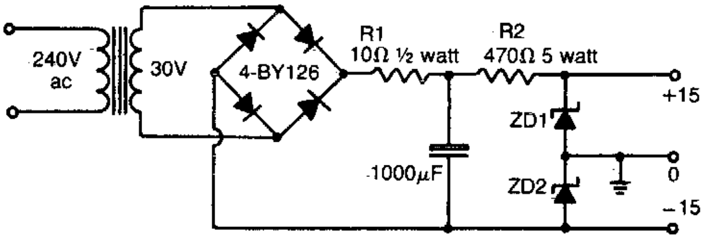
+IN
+OUT
+IN
+OUT
330Ω
A
B

Fig. 64-12

Circuit Notes

The simple zener shunt in A may not handle sufficient current if the zener available is of low wattage. A power transistor will do most of the work for the zener as shown in B. Once the zener starts conducting, a bias voltage develops across the resistor (330 Ω to 1 K), turning on the transistor. The output voltage is 0.7 V greater than the zener voltage.

DUAL POLARITY POWER SUPPLY


240V ac
30V
4-BY126
R1
10Ω ½ watt
R2
470Ω 5 watt
+15
ZD1
1000μF
ZD2
0
−15

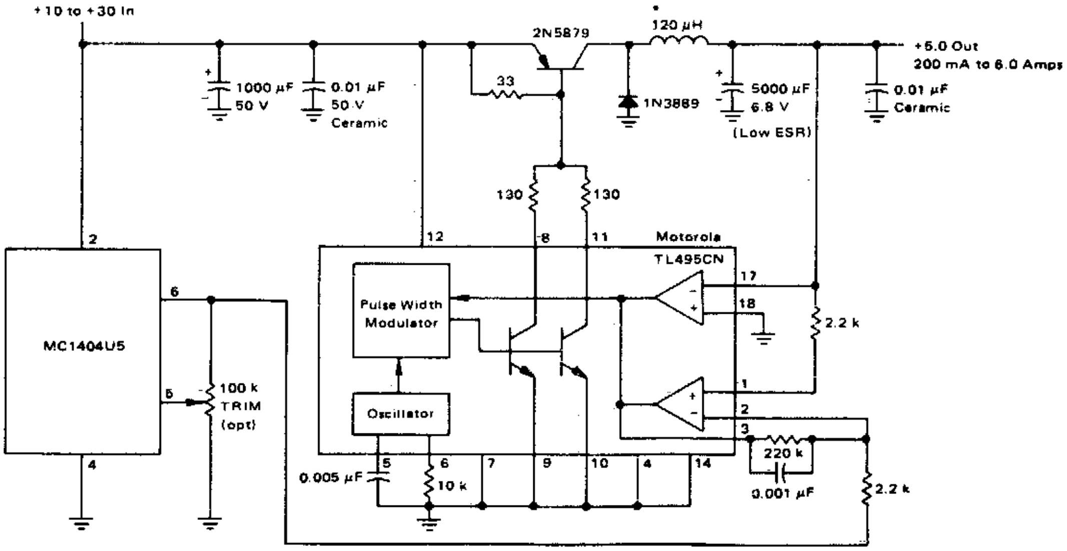
ZD 1 and ZD 2 = 15V ZENER DIODES—BZX70 C15

Fig. 64-13

Circuit Notes

This simple circuit gives a positive and negative supply from a single transformer winding and one full-wave bridge. Two zener diodes in series provide the voltage division and their centerpoint is grounded. (The filter capacitor must not be grounded via its case).

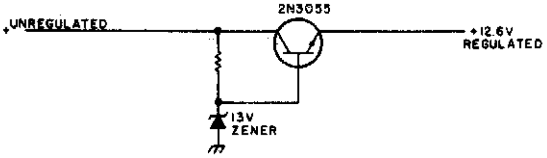
5.0 V/6.0 A 25 kHz SWITCHING REGULATOR WITH SEPARATE ULTRA-STABLE REFERENCE


+10 to +30 In
1000 μF 50 V
0.01 μF 50 V Ceramic
33
2N5879
*
120 μH
1N3889
5000 μF 6.8 V (Low ESR)
+5.0 Out
200 mA to 6.0 Amps
0.01 μF Ceramic
130
130
2
12
8
11
Motorola
TL495CN
MC1404U5
6
5
4
100 k TRIM (opt)
Pulse Width Modulator
Oscillator
17
18
2.2 k
1
2
3
220 k
0.001 μF
2.2 k
0.005 μF
5
6
10 k
7
9
10
4
14

*40 Turns #16 Wire, Arnold A-894075-2 Ferrite Core

Fig. 64-14

MOBILE VOLTAGE REGULATOR



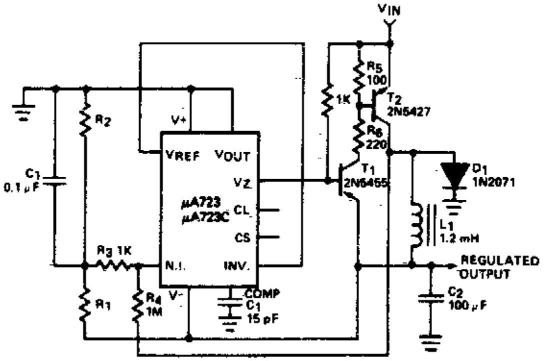
+UNREGULATED
2N3055
+12.6V
REGULATED
13V
ZENER

Fig. 64-15

Circuit Notes

This simple mobile voltage regulator circuit may save your two meter or CB transceiver if the voltage regulator fails. The 2N3055 should be heat sinked if current drawn by the rig is in excess of 2 A on transmit. This circuit will do little under normal operating conditions, but could save expensive equipment if the vehicle's electrical system loses regulation.

NEGATIVE SWITCHING REGULATOR



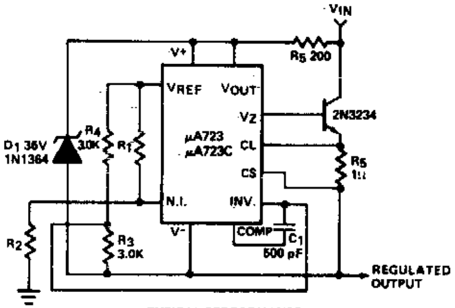
VIN
R5
100
T2
2N5427
1K
R6
220
T1
2N5455
D1
1N2071
L1
1.2 mH
REGULATED
OUTPUT
C2
100 μF
R2
V+
VREF
VOUT
VZ
μA723
μA723C
CL
CS
N.I.
INV.
C1
0.1 μF
R3 1K
R1
R4
1M
V-
COMP
C1
15 pF

TYPICAL PERFORMANCE

Regulated Output Voltage	-15V
Line Regulation (ΔV_{IN} = 20V)	8 mV
Load Regulation (ΔI_L = 2A)	6 mV

Fig. 64-16

POSITIVE FLOATING REGULATOR



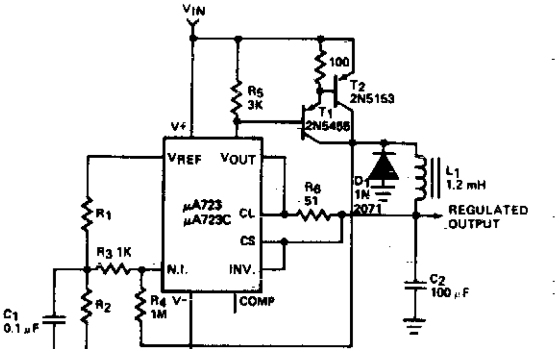
VIN
R5 200
V+
VREF
VOUT
VZ
2N3234
μA723
μA723C
CL
CS
R6
1Ω
D1 36V
1N1364
R4
30K
R1
N.I.
INV.
R2
R3
3.0K
V-
COMP
C1
500 pF
REGULATED
OUTPUT

TYPICAL PERFORMANCE

Regulated Output Voltage	+50V
Line Regulation (ΔV_{IN} = 20V)	15 mV
Load Regulation (ΔI_L = 50 mA)	20 mV

Fig. 64-18

POSITIVE SWITCHING REGULATOR



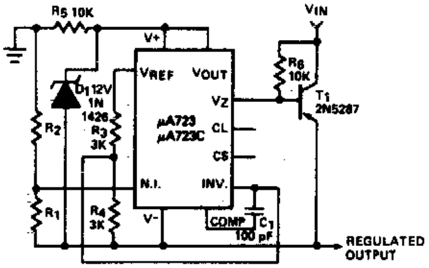
VIN
100
R5
3K
T2
2N5153
T1
2N5465
V+
VREF
VOUT
μA723
μA723C
CL
CS
N.I.
INV.
R6
51
D1
1N
2071
L1
1.2 mH
REGULATED
OUTPUT
C2
100 μF
R1
R3 1K
R2
R4
1M
V-
COMP
C1
0.1 μF

TYPICAL PERFORMANCE

Regulated Output Voltage	+5V
Line Regulation (ΔV_{IN} = 30V)	10 mV
Load Regulation (ΔI_L = 2A)	80 mV

Fig. 64-17

NEGATIVE FLOATING REGULATOR



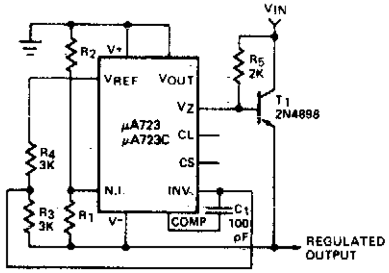
R5 10K
VIN
V+
R6
10K
T1
2N5287
VREF
VOUT
VZ
D1 12V
1N
1426
R3
3K
R2
μA723
μA723C
CL
CS
N.I.
INV.
R1
R4
3K
V-
COMP
C1
100 pF
REGULATED
OUTPUT

TYPICAL PERFORMANCE

Regulated Output Voltage	-100V
Line Regulation (ΔV_{IN} = 20V)	30 mV
Load Regulation (ΔI_L = 100 mA)	20 mV

Fig. 64-19

NEGATIVE VOLTAGE REGULATOR


VIN
R2
V+
VREF
VOUT
R5
2K
T1
2N4898
VZ
μA723
μA723C
CL
CS
R4
3K
N.I.
INV.
C1
100
pF
R3
3K
R1
V-
COMP
REGULATED
OUTPUT

TYPICAL PERFORMANCE

Regulated Ouput Voltage	-15V
Line Regulation (ΔV_{IN} = 3V)	1 mV
Load Regulation (ΔI_L = 100 mA)	2 mV

Fig. 64-20

–15 V NEGATIVE REGULATOR

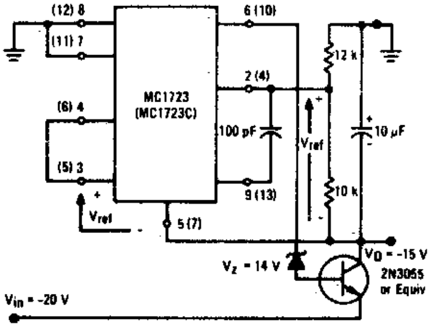

(12) 8
(11) 7
6 (10)
12 k
2 (4)
(6) 4
MC1723
(MC1723C)
100 pF
+
Vref
10 μF
(5) 3
+
Vref
-
9 (13)
10 k
5 (7)
Vz = 14 V
V0 = -15 V
2N3055
or Equiv
Vin = -20 V

Fig . 64-21

SLOW TURN-ON 15 V REGULATOR

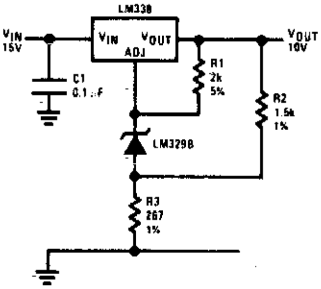

LM338
VIN
15V
VIN
VOUT
ADJ
VOUT
10V
C1
0.1 μF
R1
2k
5%
R2
1.5k
1%
LM329B
R3
267
1%

Fig. 64-22

HIGH STABILITY 10 V REGULATOR

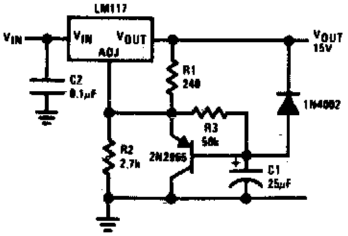

LM117
VIN
VIN
VOUT
ADJ
VOUT
15V
C2
0.1μF
R1
240
1N4002
R3
50k
R2
2.7k
2N2905
C1
25μF

Fig. 64-23

5 V/1 A SWITCHING REGULATOR

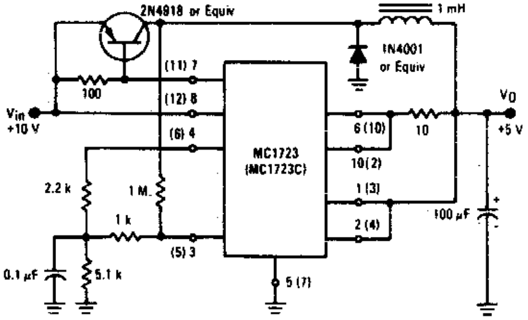

2N4918 or Equiv
1 mH
1N4001
or Equiv
(11) 7
100
(12) 8
Vin
+10 V
V0
+5 V
6 (10)
10
(6) 4
MC1723
(MC1723C)
10 (2)
2.2 k
1 M
1 (3)
1 k
100 μF
2 (4)
(5) 3
0.1 μF
5.1 k
5 (7)

Fig. 64-24

15 V/1 A REGULATOR WITH REMOTE SENSE

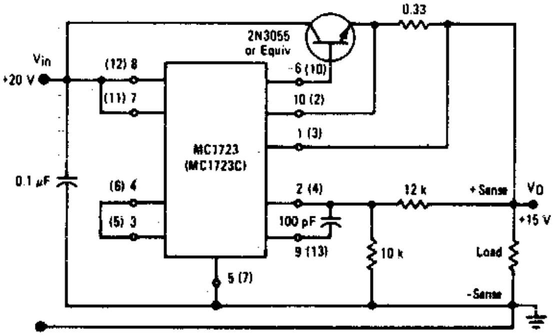
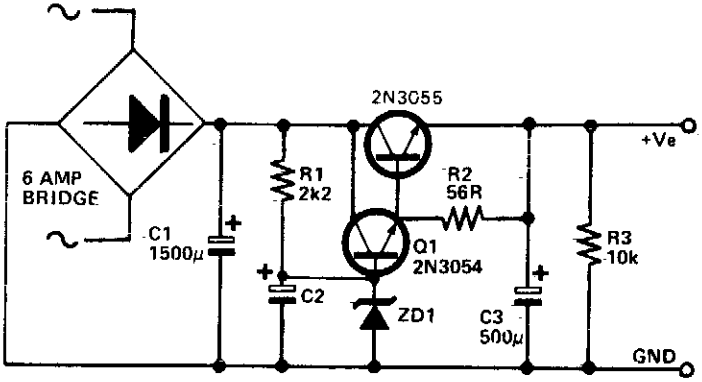

0.33
2N3055
or Equiv
Vin
+20 V
(12) 8
6 (10)
(11) 7
10 (2)
1 (3)
MC1723
(MC1723C)
0.1 μF
(6) 4
2 (4)
12 k
+ Sense
V0
+15 V
(5) 3
100 pF
9 (13)
10 k
Load
5 (7)
- Sense

Fig. 64-25

LOW RIPPLE POWER SUPPLY



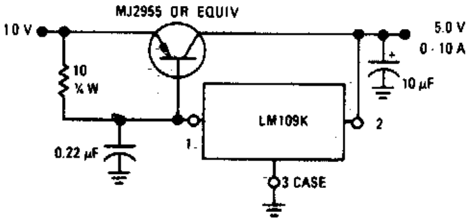
2N3055
6 AMP
BRIDGE
C1
1500μ
R1
2k2
R2
56R
Q1
2N3054
C2
ZD1
C3
500μ
R3
10k
+Ve
GND

Fig. 64-26

Circuit Notes

This circuit may be used where a high current is required with a low ripple voltage (such as in a high powered class AB amplifier when high quality reproduction is necessary). Q1, Q2, and R2 may be regarded as a power darlington transistor. ZD1 and R1 provide a reference voltage at the base of Q1. ZD1 should be chosen thus: ZD1 = $V_{out} - 1.2$. C2 can be chosen for the degree of smoothness as its value is effectively multiplied by the combined gains of Q1/Q2, if 100 μF is chosen for C2, assuming minimum hfe for Q1 and Q2, C = 100 × 15(Q1) × 25(Q2) = 37,000 μF.

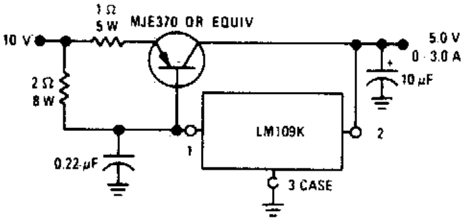
5.0 V/10 A REGULATOR



MJ2955 OR EQUIV
10 V
10
¼ W
0.22 μF
LM109K
1
2
3 CASE
5.0 V
0 - 10 A
10 μF

Fig. 64-27

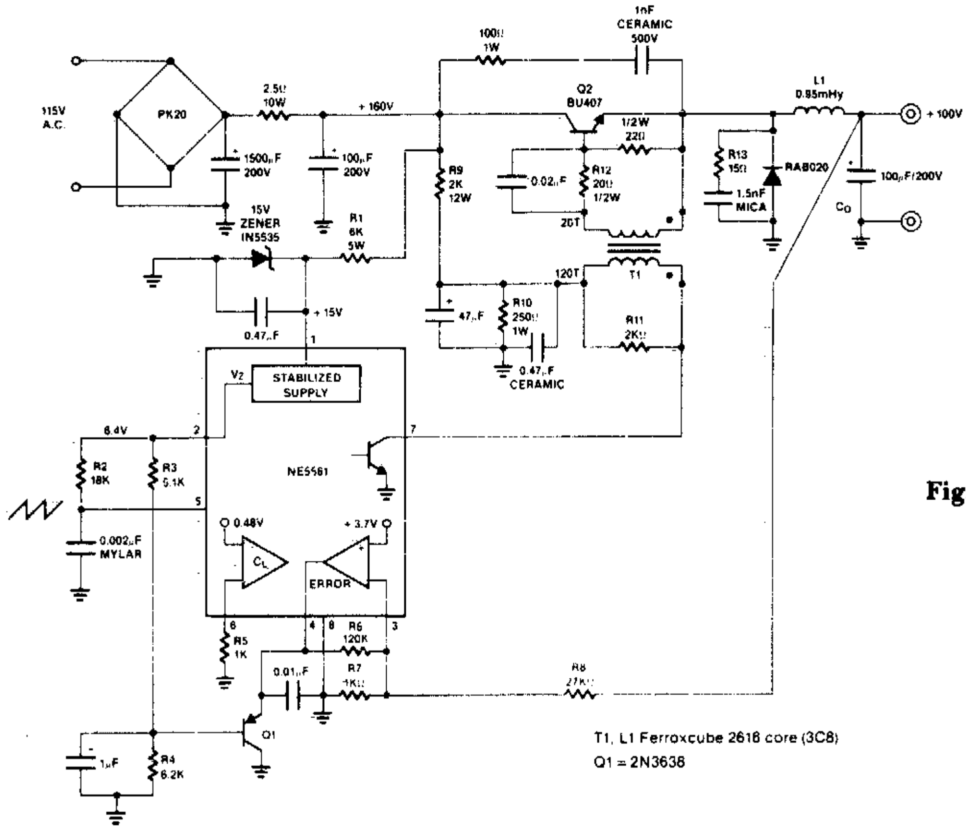
5.0 V/3.0 A REGULATOR



1 Ω
5 W
MJE370 OR EQUIV
10 V
2 Ω
8 W
0.22 μF
LM109K
1
2
3 CASE
5.0 V
0 - 3.0 A
10 μF

Fig. 64-28

100 V/0.25 A SWITCH MODE CONVERTER

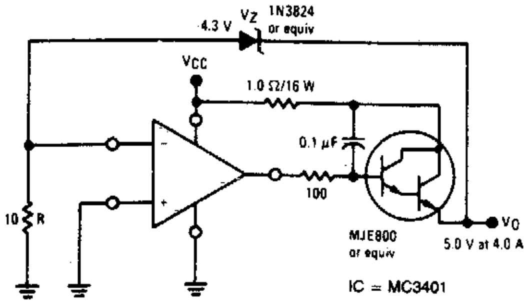


115V A.C.
PK20
2.5Ω 10W
+160V
1500μF 200V
100μF 200V
15V ZENER 1N5535
R1 6K 5W
0.47μF
+15V
1
100Ω 1W
1nF CERAMIC 500V
Q2 BU407
1/2W 22Ω
R9 2K 12W
0.02μF
R12 20Ω 1/2W
20T
120T
T1
47μF
R10 250Ω 1W
0.47μF CERAMIC
R11 2KΩ
L1 0.85mHy
+100V
R13 15Ω
1.5nF MICA
RA8020
100μF/200V
Co
VZ
STABILIZED SUPPLY
NE5561
6.4V
2
5
7
R2 18K
R3 5.1K
0.002μF MYLAR
0.48V
+3.7V
CL
ERROR
6
4
8
3
R5 1K
R6 120K
0.01μF
R7 4KΩ
R8 27KΩ
Q1
1μF
R4 6.2K

T1, L1 Ferroxcube 2616 core (3C8)
Q1 = 2N3638

Fig. 64-29

VOLTAGE REGULATOR



4.3 V
VZ
1N3824 or equiv
VCC
1.0 Ω/16 W
0.1 μF
100
−
+
10
R
MJE800 or equiv
VO
5.0 V at 4.0 A

IC = MC3401

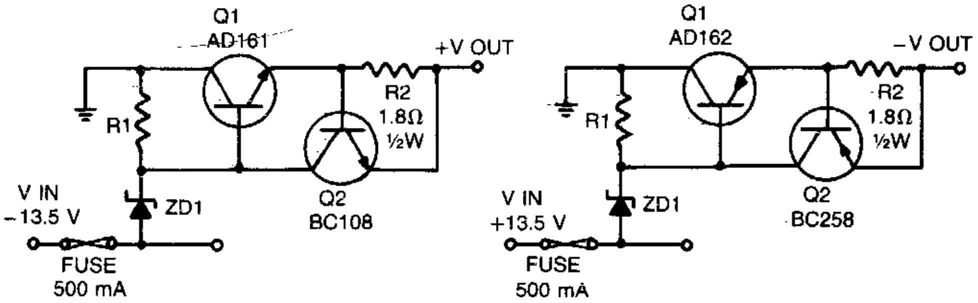
$V_O = V_Z + 0.6$ Vdc

NOTE 1: R is used to bias the zener.

NOTE 2: If the Zener TC is positive, and equal in magnitude to the negative TC of the input to the operational amplifier (≈ 2.0 mV/°C), the output is zero-TC. A 7.0-Volt Zener will give approximately zero-TC.

Fig. 64-30

LOW VOLTAGE REGULATORS WITH SHORT CIRCUIT PROTECTION


Q1
AD161
+V OUT
R2
1.8Ω
½W
R1
V IN
−13.5 V
ZD1
Q2
BC108
FUSE
500 mA
Q1
AD162
−V OUT
R2
1.8Ω
½W
R1
V IN
+13.5 V
ZD1
Q2
BC258
FUSE
500 mA

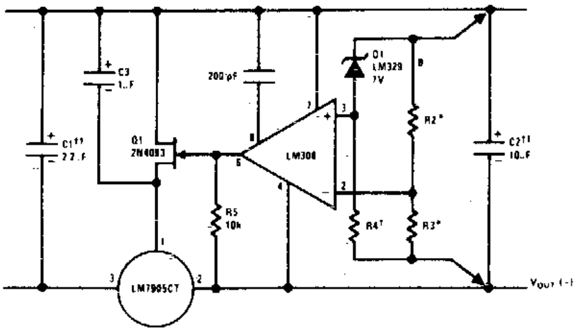
VOLTAGE	ZD1 400mW	R1
6V	6V2	680Ω
7.5V	7V5	390Ω
9V	9V1	220Ω

Fig. 64-31

Circuit Notes

These short-circuit protected regulators give 6, 7.5, and 9 V from an automobile battery supply of 13.5 V nominal; however, they will function just as well if connected to a smoothed dc output from a transformer/rectifier circuit. Two types are shown for both positive and negative ground systems. The power transistors can be mounted on the heatsink without a mica insulating spacer thus allowing for greater cooling efficiency. Both circuits are protected against overload or short-circuits. The current cannot exceed 330 mA. Under normal operating conditions the voltage across R2 does not rise above the 500 mV necessary to turn Q2 on and the circuit behaves as if there was only Q1 present. If excessive current is drawn, Q2 turns on and cuts off Q1, protecting the regulating transistor. The table gives the values of R1 for different zener voltages.

HIGH STABILITY 1 A REGULATOR


C3
1µF
200pF
D1
LM329
7V
8
C1††
2.2µF
Q1
2N4093
7
3
R2*
C2††
10µF
6
LM308
4
2
R5
10k
R4†
R3*
1
3
LM7905CT
2
V_{OUT} (−)

Load and line regulation < 0.01% temperature stability ≤ 0.2%

† Determines Zener current

†† Solid tantalum

* Select resistors to set output voltage. 2 ppm/°C tracking suggested

Fig. 64-32

100 V/0.25 A SWITCH MODE CONVERTER

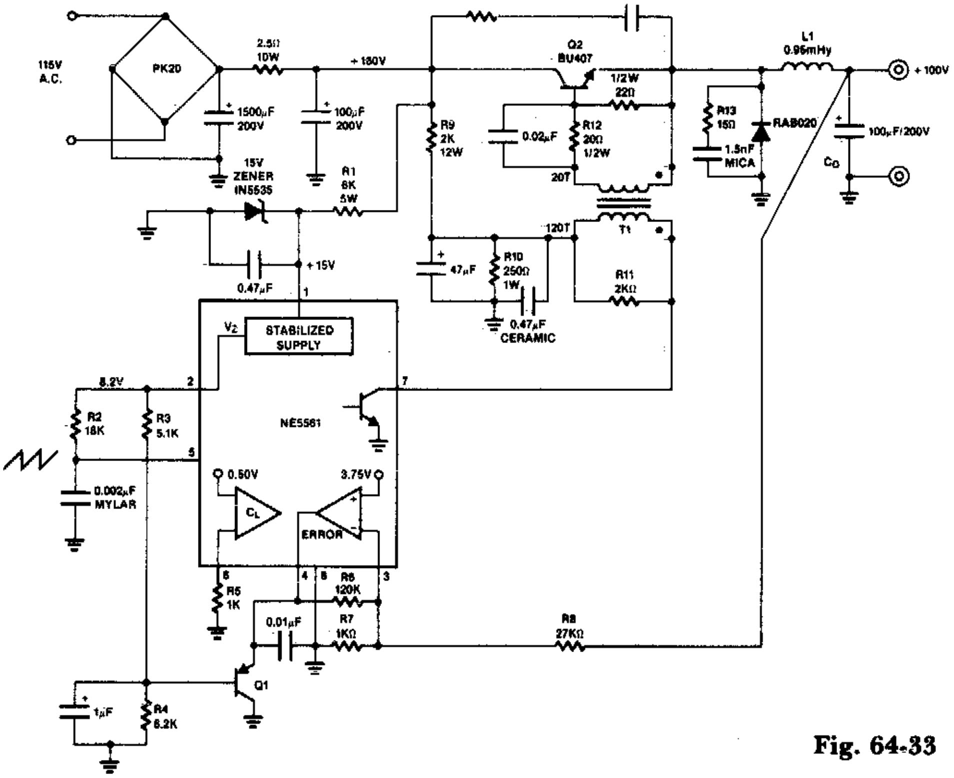

115V A.C.
PK20
2.5Ω 10W
+180V
1500μF 200V
100μF 200V
15V ZENER IN5535
R1 6K 5W
0.47μF
+15V
Q2 BU407
1/2W 22Ω
0.02μF
R12 20Ω 1/2W
R9 2K 12W
20T
120T
T1
47μF
R10 250Ω 1W
0.47μF CERAMIC
R11 2KΩ
L1 0.96mHy
+100V
R13 15Ω
1.5nF MICA
RAB020
100μF/200V
C_O
STABILIZED SUPPLY
V_Z
NE5561
8.2V
R2 18K
R3 5.1K
0.002μF MYLAR
0.50V
3.75V
C_L
ERROR
1
2
3
4
5
6
7
8
R5 1K
R6 120K
R7 1KΩ
R8 27KΩ
0.01μF
1μF
R4 6.2K
Q1

Fig. 64-33

65

Power Supplies (Variable)

The sources of the following circuits are contained in the Sources section beginning on page 730. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

Dual Output Bench Power Supply
Power Supply with Adjustable Current Limit and Output Voltage
Adjustable Output Regulator
10 mA Negative-Voltage from a Positive Source
Regulated Voltage Divider
Variable Zener Diode
12 V To 9, 7.5 or 6 V Converter
5 A Constant Voltage/Constant Current Regulator
Power Pack for Battery-Powered Calculators, Radios, or Cassette Players
Precision High Voltage Regulator
Remote Shutdown Regulator with Current Limiting
0 to 22 V Regulator
0 to 30 V Regulator
10 A Regulator
Adjustable Regulator 0-10 V at 3 A
High Voltage Regulator
Low Voltage Regulator
Simple Split Power Supply
Adjustable Output Regulator
Multiple Output Switching Regulator for Use with MPUs
6.0 A Variable Output Switching Regulator

DUAL OUTPUT BENCH POWER SUPPLY

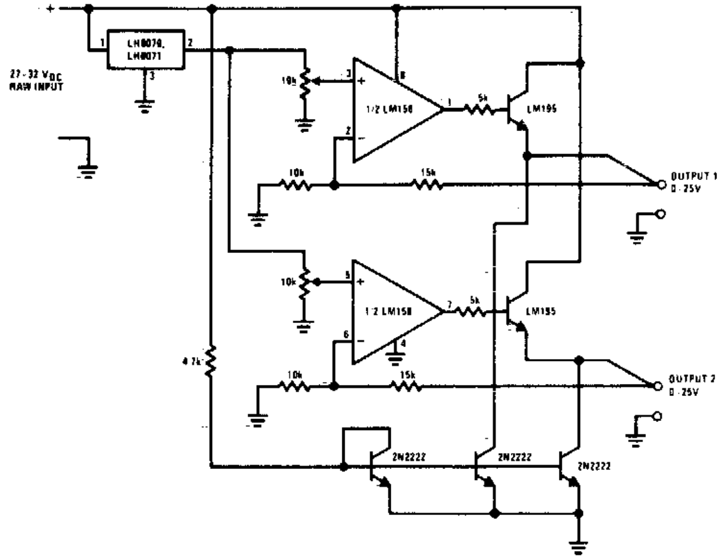
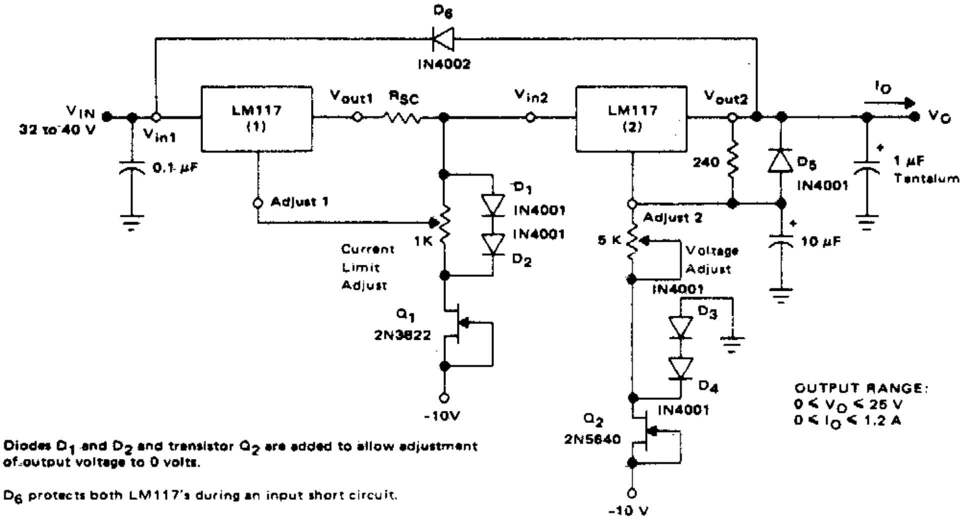

+
LH0070,
LH0071
1
2
3
27 - 32 V$_{DC}$
RAW INPUT
10k
3
+
8
1/2 LM158
2
−
1
5k
LM195
10k
15k
OUTPUT 1
0 - 25V
10k
5
+
1/2 LM158
6
−
4
7
5k
LM195
4.7k
10k
15k
OUTPUT 2
0 - 25V
2N2222
2N2222
2N2222

Fig. 65-1

POWER SUPPLY WITH ADJUSTABLE CURRENT LIMIT AND OUTPUT VOLTAGE


D$_6$
IN4002
V$_{IN}$
32 to 40 V
V$_{in1}$
0.1 μF
LM117
(1)
V$_{out1}$
R$_{SC}$
Adjust 1
D$_1$
IN4001
IN4001
D$_2$
1K
Current
Limit
Adjust
Q$_1$
2N3822
-10V
V$_{in2}$
LM117
(2)
V$_{out2}$
I$_O$
V$_O$
240
D$_5$
IN4001
1 μF
Tantalum
Adjust 2
5 K
Voltage
Adjust
10 μF
IN4001
D$_3$
D$_4$
IN4001
Q$_2$
2N5640
-10 V

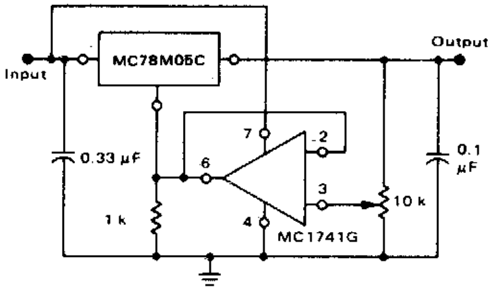
Diodes D_1 and D_2 and transistor Q_2 are added to allow adjustment of output voltage to 0 volts.

D_6 protects both LM117's during an input short circuit.

OUTPUT RANGE:
$0 \leq V_O \leq 25$ V
$0 \leq I_O \leq 1.2$ A

Fig. 65-2

ADJUSTABLE OUTPUT REGULATOR


Input
MC78M05C
Output
7
2
6
3
4
0.33 μF
1 k
10 k
0.1 μF
MC1741G

V_O. 7.0 V to 20 V
V_{IN} $V_O \geqslant 2.0$ V

Fig. 65-3

Circuit Notes

The addition of an operational amplifier allows adjustment to higher or intermediate values while retaining regulation characteristics. The minimum voltage obtainable with this arrangement is 2.0 volts greater than the regulator voltage.

RF PROBE FOR VTVM

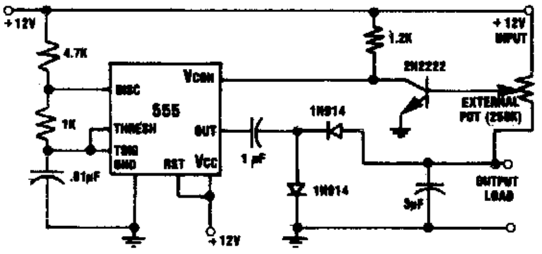

+12V
4.7K
1K
.01μF
DISC
555
THRESH
TRIG
GND
RST
Vcc
VCON
OUT
+12V
1 μF
1N914
1N914
1.2K
2N2222
+12V INPUT
EXTERNAL POT (250K)
3μF
OUTPUT LOAD

Fig. 65-4

Circuit Notes

This circuit combines a 555 timer with a 2N2222 transistor and an external potentiometer. The pot adjusts the output voltage to the desired value. To regulate the output voltage, the 2N2222 varies the control voltage of the 555 IC, increasing or decreasing the pulse repetition rate. A 1.2 K resistor is used as a collector load. The transistor base is driven from the external pot. If the output voltage becomes less negative, the control voltage moves closer to ground, causing the repetition rate of the 555 to increase, which, in turn, causes the 3 μF capacitor to charge more frequently. Output voltage for the circuit is 0 to 10 V, adjusted by the external pot. Output regulation is less than five percent for 0 to 10 mA and less than .05 percent for 0 to 0.2 mA.

REGULATED VOLTAGE DIVIDER

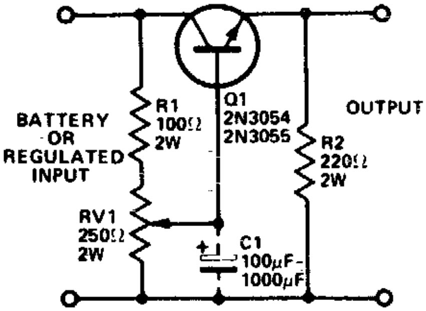


Fig. 65-5

Circuit Notes

ICs requiring 3.6 or 6 volts can be run from a battery or fixed regulated supply of a higher voltage by using the circuit shown. The transistor should be mounted on a heatsink as considerable power will be dissipated by its collector. Additional filtering can be obtained by fitting a capacitor (C1) as shown. The capacitance is effectively multiplied by the gain of the transistor. A ripple of 200 mV (peak to peak) at the input can be reduced to 2 mV in this fashion. Maximum output current depends on the supply rating and transistor type (with heatsink) used.

VARIABLE ZENER DIODE

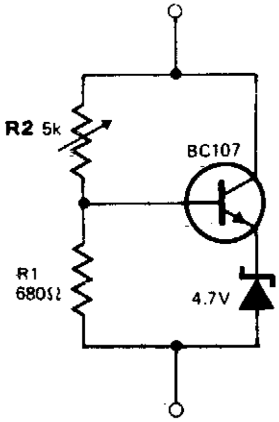
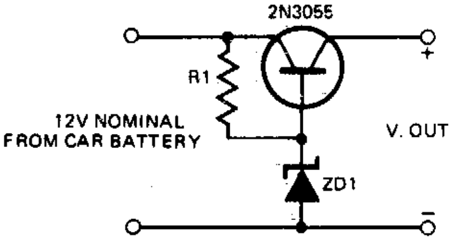


Fig. 65-6

Circuit Notes

The circuit behaves like a zener diode over a large range of voltages. The current passing through the voltage divider R1-R2 is substantially larger than the transistor base current and is in the region of 8 mA. The stabilizing voltage is adjustable over the range 5-45 V by changing the value of R2. The total current drawn by the circuit is variable over the range 15 mA to 50 mA. This value is determined by the maximum dissipation of the zener diode. In the case of a 250 mW device, this is of the order of 50 mA.

12 V TO 9, 7.5 or 6 V CONVERTER


2N3055
R1
12V NOMINAL
FROM CAR BATTERY
+
V. OUT
ZD1
−

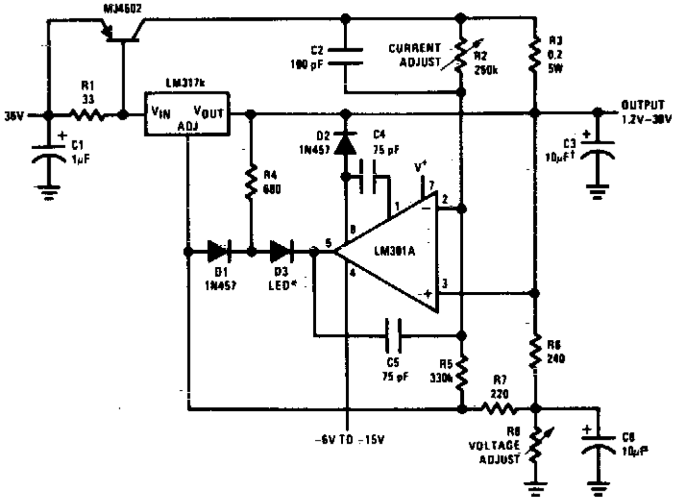
OUTPUT VOLTAGE	9	7.5	6
R1 (½ WATT)	180Ω	270Ω	330Ω
ZENER DIODE (250 mW)	10V	8V1	6V6

Fig. 65-7

Circuit Notes

This circuit enables transistorized items such as radio, cassettes, and other electrical devices to be operated from a car's electrical supply. The table gives values for resistors and specified diode types for different voltage. Should more than one voltage be required a switching arrangement could be incorporated. For high currents, the transistor should be mounted on a heatsink.

5 A CONSTANT VOLTAGE/CONSTANT CURRENT REGULATOR

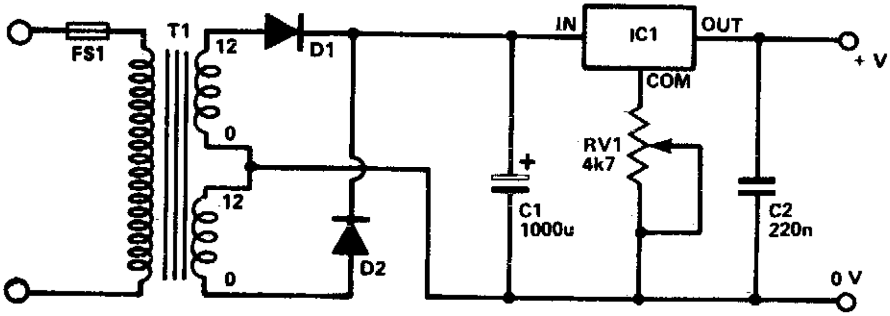

MJ4502
R1
33
35V
C1
1μF
LM317k
V_{IN}
V_{OUT}
ADJ
C2
100 pF
CURRENT ADJUST
R2
250k
R3
0.2
5W
OUTPUT
1.2V–30V
C3
10μF†
D2
1N457
C4
75 pF
R4
680
V+
LM301A
D1
1N457
D3
LED*
C5
75 pF
R5
330k
R6
240
R7
220
R8
VOLTAGE ADJUST
C6
10μF
−6V TO −15V

†Solid tantalum

*Lights in constant current mode

Fig. 65-8

POWER PACK FOR BATTERY-POWERED CALCULATORS, RADIOS, OR CASSETTE PLAYERS


FS1
T1
12
0
12
0
D1
D2
C1
1000u
IN
IC1
OUT
COM
RV1
4k7
C2
220n
+ V
0 V

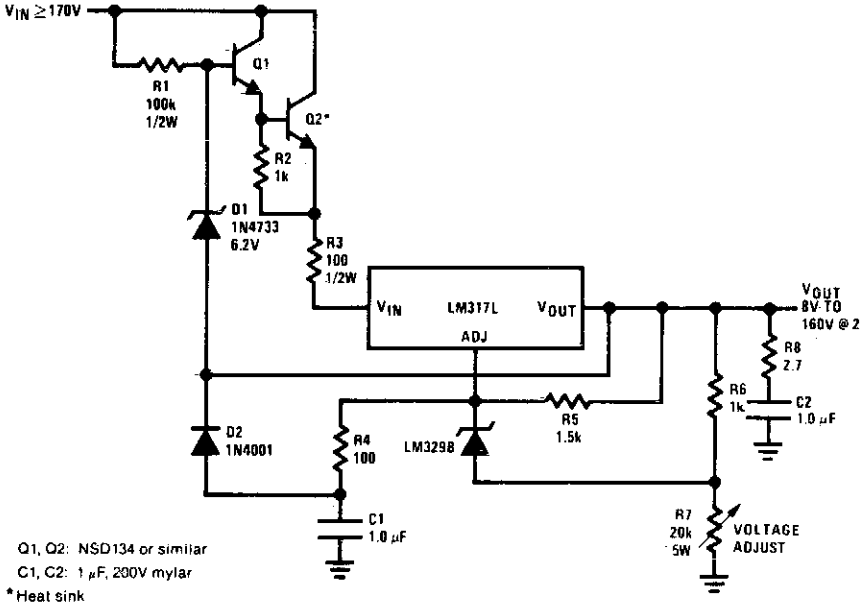
NOTES:
IC1 IS 7805
D1,2 ARE 1N4001

Fig. 65-9

Circuit Notes

This circuit gives a regulated output of between 5 V and 15 Vdc, adjusted and set by a preset resistor. Current output up to about 350 mA. An integrated circuit regulates the output voltage and although this IC (the 7805) is normally used in a fixed-voltage (5 Vdc) supply it is for a variable output voltage.

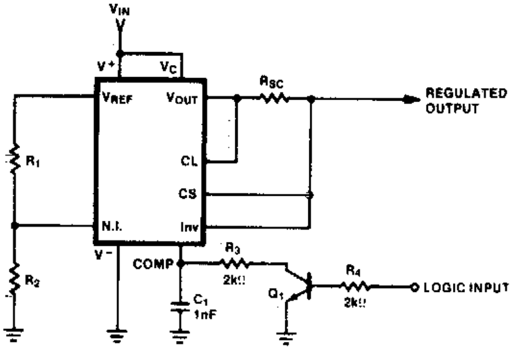
PRECISION HIGH VOLTAGE REGULATOR


$V_{IN} \geq 170V$
R1
100k
1/2W
Q1
Q2*
R2
1k
D1
1N4733
6.2V
R3
100
1/2W
V_{IN}
LM317L
V_{OUT}
ADJ
V_{OUT}
8V TO
160V @ 25 mA
R8
2.7
R6
1k
C2
1.0 μF
R5
1.5k
D2
1N4001
R4
100
LM329B
C1
1.0 μF
R7
20k
5W
VOLTAGE
ADJUST

Q1, Q2: NSD134 or similar
C1, C2: 1 μF, 200V mylar
*Heat sink

Fig. 65-10

REMOTE SHUTDOWN REGULATOR WITH CURRENT LIMITING (V_{out} = 2 TO 7 V)



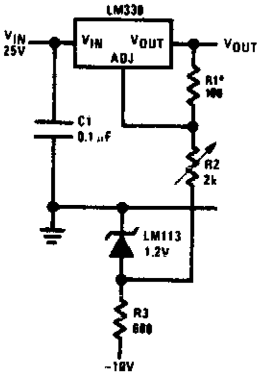
VIN
V+
VC
VREF
VOUT
RSC
REGULATED OUTPUT
R1
CL
CS
N.I.
Inv
V−
COMP
R3
2kΩ
R2
C1
1nF
Q1
R4
2kΩ
LOGIC INPUT

IC = μA723

$$V_{OUT} = \left[V_{REF} \times \frac{R_2}{R_1 + R_2}\right]$$

Fig. 65-11

0 TO 22 V REGULATOR

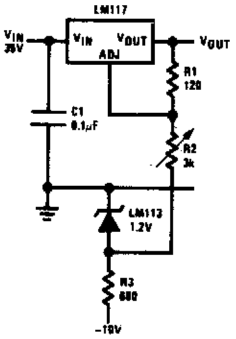


LM338
VIN
25V
VIN
VOUT
ADJ
VOUT
R1*
100
C1
0.1 μF
R2
2k
LM113
1.2V
R3
680
−10V

*R1=240Ω, R2 = 5k for LM138 and LM238

Fig. 65-12

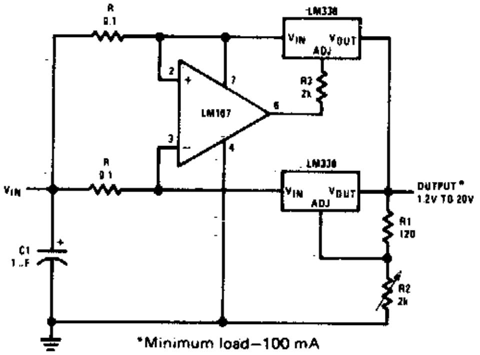
0 TO 30 V REGULATOR



LM117
VIN
35V
VIN
VOUT
ADJ
VOUT
R1
120
C1
0.1μF
R2
3k
LM113
1.2V
R3
680
−10V

Fig. 65-13

10 A REGULATOR

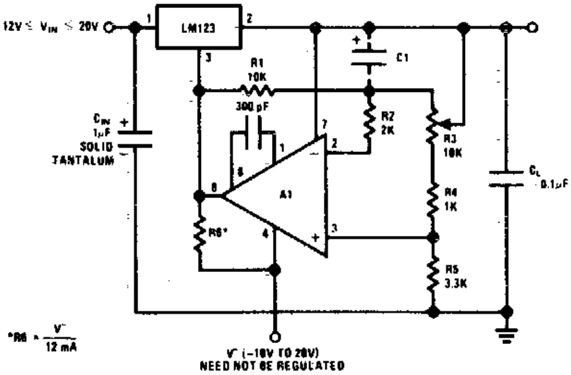


R
0.1
LM338
VIN
VOUT
ADJ
2
+
7
R2
2k
LM107
6
3
−
4
R
0.1
LM338
VIN
VIN
VOUT
ADJ
OUTPUT*
1.2V TO 20V
R1
120
C1
1 μF
+
R2
2k

*Minimum load—100 mA

Fig. 65-14

ADJUSTABLE REGULATOR 0-10 V AT 3 A


12V ≤ VIN ≤ 20V
LM123
1
2
3
R1
10K
300 pF
C1
R2
2K
R3
10K
R4
1K
R5
3.3K
CIN
1μF
SOLID
TANTALUM
CL
0.1μF
A1
R6*
V⁻ (−10V TO 20V)
NEED NOT BE REGULATED

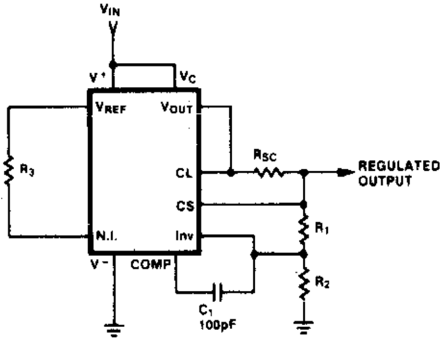
$^*R6 = \frac{V^-}{12\ mA}$

A₁ – LM101A

C₁ – 2μF OPTIONAL – IMPROVES RIPPLE REJECTION, NOISE, AND TRANSIENT RESPONSE

Fig. 65-15

HIGH VOLTAGE REGULATOR (Vout = + 7 V TO 37 V)


VIN
V+
VC
VREF
VOUT
R3
CL
CS
N.I.
Inv
V−
COMP
RSC
REGULATED OUTPUT
R1
R2
C1
100pF

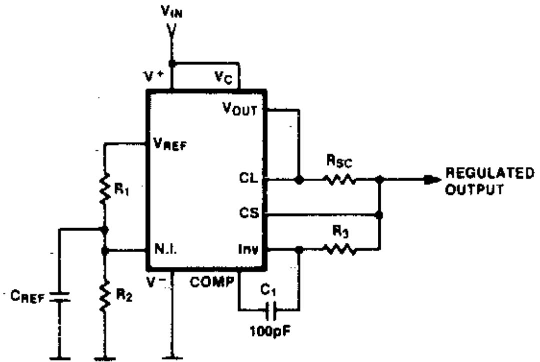
$V_{OUT} = \left[V_{REF} \times \frac{R_1 + R_2}{R_2}\right]$

$R_3 = \frac{R_1 R_2}{R_1 + R_2}$ for minimum temperature drift

R3 may be eliminated for minimum component count

Fig. 65-16

LOW VOLTAGE REGULATOR (Vout = 2 TO 7 V)


VIN
V+
VC
VOUT
VREF
R1
CL
CS
N.I.
Inv
V−
COMP
RSC
REGULATED OUTPUT
R3
CREF
R2
C1
100pF

$V_{OUT} = \left[V_{REF} \times \frac{R_2}{R_1 + R_2}\right]$

$R_3 = \frac{R_1 R_2}{R_1 + R_2}$ for minimum temperature drift

Fig. 65-17

SIMPLE SPLIT POWER SUPPLY

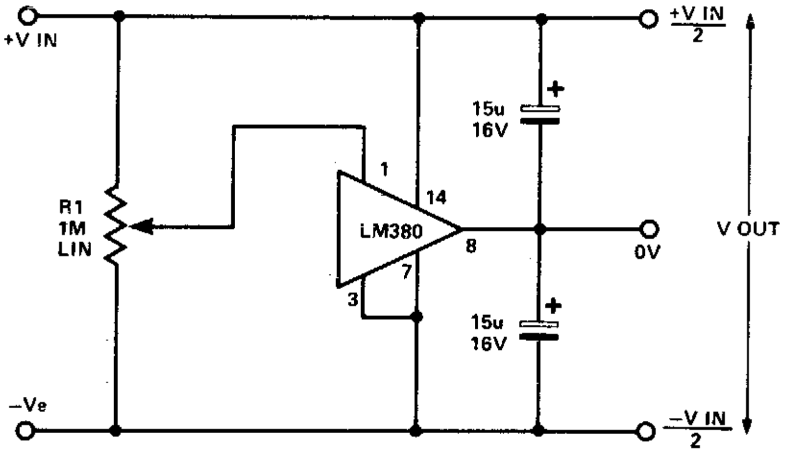

+V IN
+V IN / 2
15u
16V
R1
1M
LIN
1
14
LM380
8
7
3
V OUT
0V
15u
16V
–Ve
–V IN / 2

Fig. 65-18

Circuit Notes

This circuit utilizes the quasi-complementary output stage of the popular LM380 audio power IC. The device is internally biased so that with no input the output is held midway between the supply rails. R1, which should be initially set to mid-travel, is used to nullify any inbalance in the output. Regulation of V_{out} depends upon the circuit feeding the LM380, but positive and negative outputs will track accurately irrespective of input regulation and unbalanced loads. The free-air dissipation is a little over 1 watt, and so extra cooling may be required. The device is fully protected and will go into thermal shutdown if its rated dissipation is exceeded. Current limiting occurs if the output current exceeds 1.3 A. The input voltage should not exceed 20 V.

ADJUSTABLE OUTPUT REGULATOR

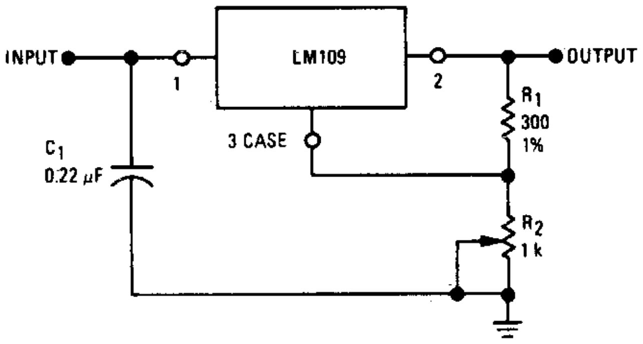
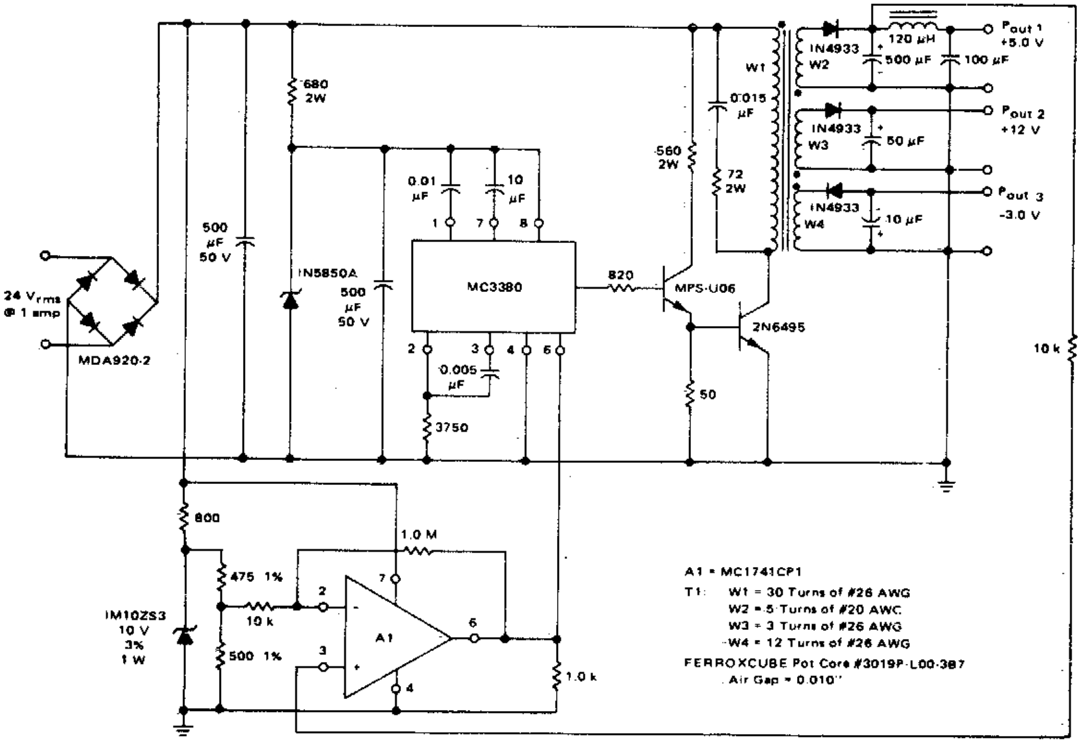

INPUT
1
LM109
2
OUTPUT
R1
300
1%
C1
0.22 μF
3 CASE
R2
1 k

Fig. 65-19

MULTIPLE OUTPUT SWITCHING REGULATOR FOR USE WITH MPUs


24 Vrms
@ 1 amp
MDA920-2
500 µF 50 V
680 2W
1N5850A
500 µF 50 V
0.01 µF
10 µF
1
7
8
MC3380
2
3
4
6
0.005 µF
3750
820
MPS-U06
2N6495
50
560 2W
0.015 µF
72 2W
W1
1N4933
W2
120 µH
500 µF
100 µF
Pout 1 +5.0 V
1N4933
W3
50 µF
Pout 2 +12 V
1N4933
W4
10 µF
Pout 3 -3.0 V
10 k
800
1.0 M
475 1%
10 k
500 1%
1M10ZS3 10 V 3% 1 W
2
3
7
6
4
A1
1.0 k

A1 = MC1741CP1

T1: W1 = 30 Turns of #26 AWG
W2 = 5 Turns of #20 AWG
W3 = 3 Turns of #26 AWG
W4 = 12 Turns of #26 AWG

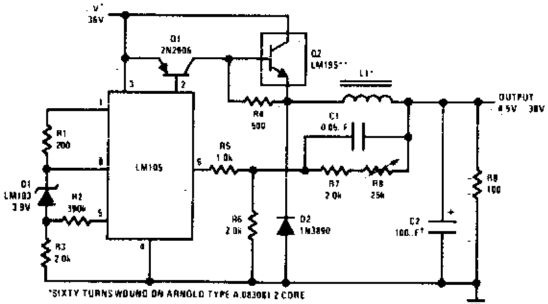
FERROXCUBE Pot Core #3019P-L00-387
Air Gap = 0.010″

A1 = MC3380P

TYPICAL PERFORMANCE
P_{out1} = 4 Watts
(V_O = 5 V ± 5%)
5 V Ripple Component = 50 mV
(120 Hz + 20 kHz)
P_{out2} = 600 mW
(V_O = 12 V ± 10%)
P_{out3} = 3 mW
(V_O = -3 V ± 10%)

Fig. 65-20

6.0 A VARIABLE OUTPUT SWITCHING REGULATOR


V+ 36V
Q1 2N2906
Q2 LM195**
3
2
1
8
5
4
6
LM105
L1*
R1 200
D1 LM103 3.9V
R2 390k
R3 2.0k
R4 500
C1 0.05 µF
R5 1.0k
R6 2.0k
R7 2.0k
R8 25k
D2 1N3890
C2 100 µF†
R9 100
OUTPUT 4.5V-30V

*SIXTY TURNS WOUND ON ARNOLD TYPE A.083081 2 CORE
**FOUR DEVICES IN PARALLEL
†SOLID TANTALUM

Fig. 65-21

66

Power Supply Protection Circuits

The sources of the following circuits are contained in the Sources section beginning on page 730. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

Electronic Crowbar for Ac or Dc Lines
Power Protection Circuit
Simple Crowbar
Overvoltage Protection with Automatic Reset
Overvoltage Protection for Logic
Fast Acting Power Supply Protection
5 V Crowbar

ELECTRONIC CROWBAR FOR AC OR DC LINES

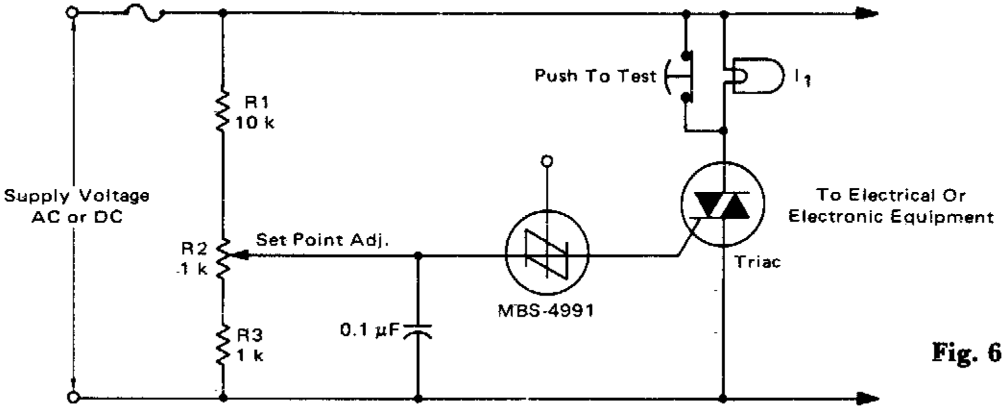

Push To Test
I1
R1
10 k
Supply Voltage
AC or DC
R2
1 k
Set Point Adj.
MBS-4991
Triac
To Electrical Or
Electronic Equipment
R3
1 k
0.1 μF

Fig. 66-1

Circuit Notes

For positive protection of electrical or electronic equipment, use this against excessive supply voltage. Due to improper switching, wiring, short circuits, or failure of regulators, an electronic crowbar circuit can quickly place a short circuit across the power lines, thereby dropping the voltage across the protected device to near zero and blowing a fuse. The triac and SBS are both bilateral devices, the circuit is equally useful on ac or dc supply lines. With the values shown for R1, R2, and R3, the crowbar operating point can be adjusted over the range of 60 to 120 volts dc or 42 to 84 volts ac. The resistor values can be changed to cover a different range of supply voltages. The voltage rating of the triac must be greater than the highest operating point as set by R2. I1 is a low power incandescent lamp with a voltage rating equal to the supply voltage. It may be used to check the set point and operation of the unit by opening the test switch and adjusting the input or set point to fire the SBS. An alarm unit such as the Mallory Sonalert may be connected across the fuse to provide an audible indication of crowbar action. (This circuit may not act on short, infrequent power line transients).

POWER PROTECTION CIRCUIT

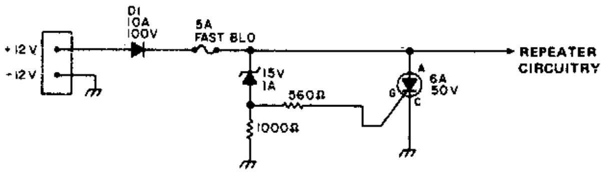

D1
10A
100V
5A
FAST BLO
+12V
-12V
15V
1A
560Ω
1000Ω
A
G
C
6A
50V
REPEATER
CIRCUITRY

Fig. 66-2

Circuit Notes

To safeguard portable, emergency power repeaters from reverse or excessive voltage, D1 prevents incorrect polarity damage, and zener voltage determines the maximum voltage that will reach the rest of the circuitry. Use fast blowing fuse rated greater than the SCR current rating.

SIMPLE CROWBAR

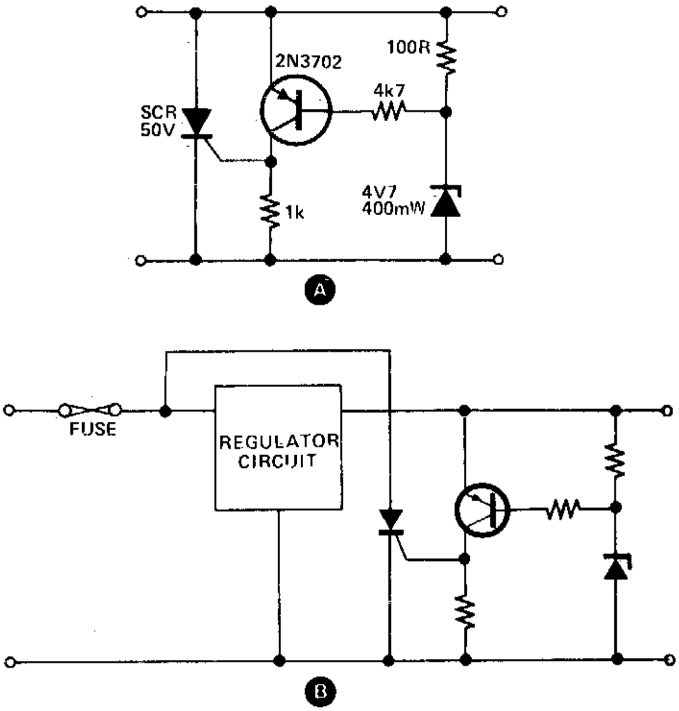
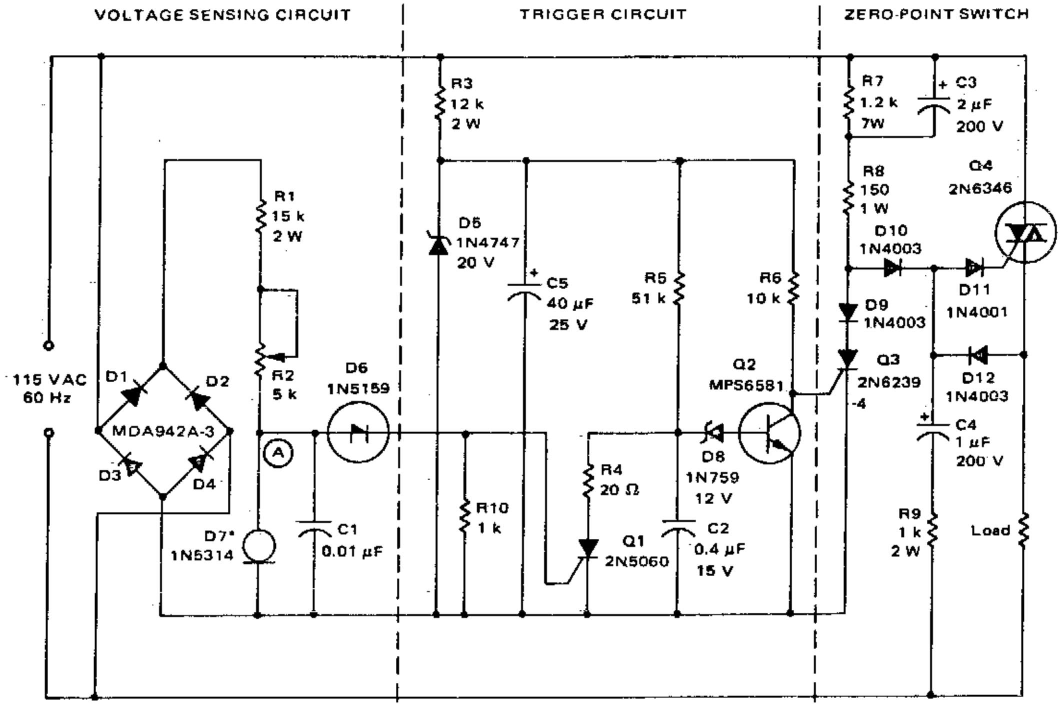


Fig. 66-3

Circuit Notes

These circuits provide overvoltage protection in case of voltage regulator failure or application of an external voltage. Intended to be used with a supply offering some form of short circuit protection, either foldback, current limiting, or a simple fuse. The most likely application is a 5 V logic supply, since TTL is easily damaged by excess voltage. The values chosen in A are for a 5 V supply, although any supply up to about 25 V can be protected by simply choosing the appropriate zener diode. When the supply voltage exceeds the zener voltage +0.7 V, the transistor turns on and fires the thyristor. This shorts out the supply, and prevents the voltage rising any further. In the case of a supply with only fuse protection, it is better to connect the thyristor the regulator circuit when the crowbar operates. The thyristor should have a current rating about twice the expected short circuit current and a maximum voltage greater than the supply voltage. The circuit can be reset by either switching off the supply, or by breaking the thyristor circuit with a switch.

OVERVOLTAGE PROTECTION WITH AUTOMATIC RESET


VOLTAGE SENSING CIRCUIT
TRIGGER CIRCUIT
ZERO-POINT SWITCH
115 VAC
60 Hz
D1
D2
D3
D4
MDA942A-3
R1
15 k
2 W
R2
5 k
A
D7*
1N5314
C1
0.01 μF
D6
1N5159
R3
12 k
2 W
D5
1N4747
20 V
C5
40 μF
25 V
R10
1 k
R4
20 Ω
Q1
2N5060
R5
51 k
R6
10 k
C2
0.4 μF
15 V
D8
1N759
12 V
Q2
MPS6581
R7
1.2 k
7W
C3
2 μF
200 V
R8
150
1 W
Q4
2N6346
D10
1N4003
D11
1N4001
D9
1N4003
Q3
2N6239
-4
D12
1N4003
C4
1 μF
200 V
R9
1 k
2 W
Load

*Two Diodes in Parallel

Fig. 66-4

OVERVOLTAGE PROTECTION FOR LOGIC

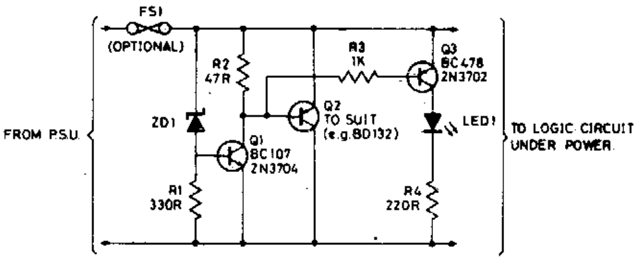

FS1
(OPTIONAL)
R2
47R
R3
1K
Q3
BC478
2N3702
ZD1
Q2
TO SUIT
(e.g.BD132)
LED1
Q1
BC107
2N3704
R1
330R
R4
220R
FROM P.S.U.
TO LOGIC CIRCUIT UNDER POWER

Fig. 66-5

Circuit Notes

Zener diode ZD1 senses the supply, and should the supply rise above 6 V, Q1 will turn on. In turn, Q2 conducts clamping the rail. Subsequent events depend on the source supply. It will either shut down, go into current limit or blow its supply fuse. None of these will damage the TTL chips. The rating of Q2 depends on the source supply, and whether it will be required to operate continuously in the event of failure. Its current rating has to be in excess of the source supply.

FAST ACTING POWER SUPPLY PROTECTION

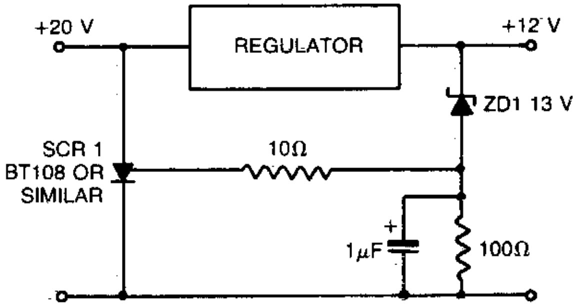

+20 V
REGULATOR
+12 V
ZD1 13 V
SCR 1
BT108 OR
SIMILAR
10Ω
+
1μF
100Ω

Fig. 66-6

Circuit Notes

When using a regulated power supply to reduce a supply voltage, there is always the danger that component failure in the power supply might lead to a severe overvoltage condition across the load. To cope with overvoltage situations, the circuit is designed to protect the load under overvoltage conditions. Component values given are for a 20 V supply with regulated output at 12 V. The zener diode can be changed according to whatever voltage is to be the maximum. If the voltage at the regulator output rises to 13 V or above, the zener diode breaks down and triggers the thyristor which shorts out the supply line and blows the main fuse.

5 V CROWBAR

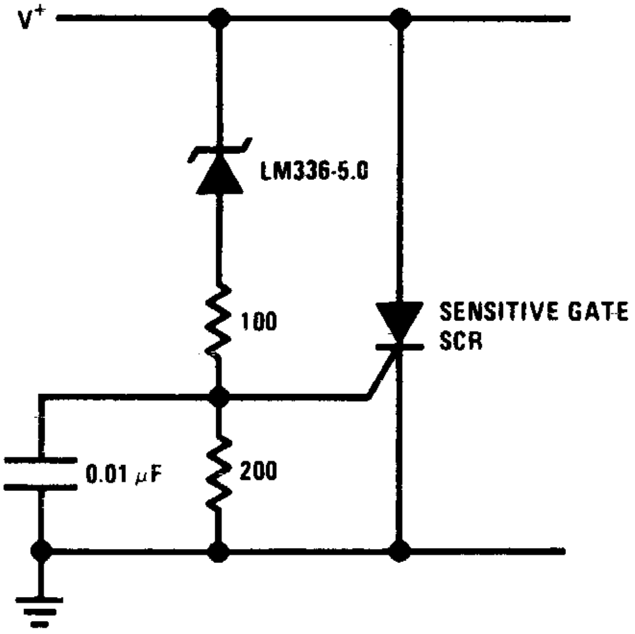

V+
LM336-5.0
100
SENSITIVE GATE
SCR
0.01 μF
200

Fig. 66-7

67

Probes

The sources of the following circuits are contained in the Sources section beginning on page 730. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

Logic Probe Yields Three Discrete States
Signal Injector/Tracer
Injector/Tracer
CMOS Logic Probe
RF Probe for VOM
100 K Megohm DC Probe
Audible TTL Probe
Logic Probe
Logic Test Probe with Memory
Logic Probe
Simple Logic Probe
Audio-RF Signal Tracer Probe
TTL Logic Tester

LOGIC PROBE YIELDS THREE DISCRETE STATES

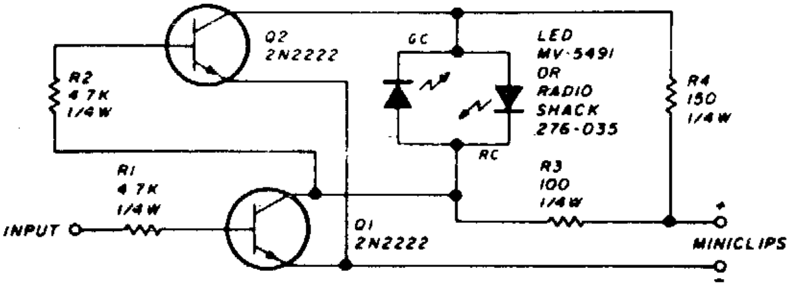
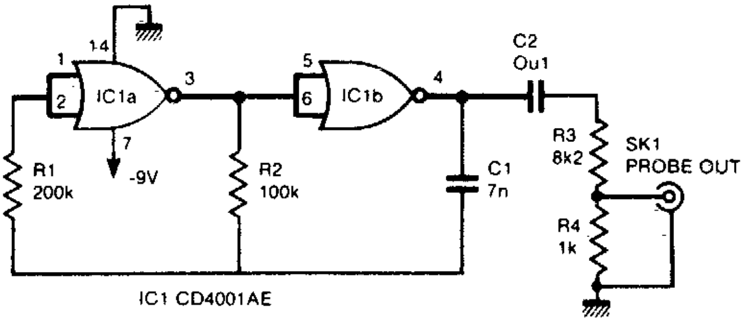

Q2
2N2222
GC
LED
MV-5491
OR
RADIO
SHACK
276-035
R2
4.7K
1/4 W
R4
150
1/4 W
RC
R1
4.7K
1/4 W
R3
100
1/4 W
INPUT
Q1
2N2222
+
MINICLIPS
-

Fig. 67-1

Circuit Notes

The circuit uses a dual LED. When power is applied to the probe through the power leads, and the input is touched to a low level or ground, Q1 is cut off. This will cause Q2 to conduct since the base is positive with respect to the emitter. With Q1 cut off and Q2 conducting, the green diode of the dual LED will be forward biased, yielding a green output. Touching the probe tip to a high level will cause Q1 and Q2 to complement, and the red diode will be forward biased, yielding a red output from the LED. An alternating signal will cause alternating conduction of the red and green diodes and will yield an indication approximately amber. In this manner, both static and dynamic signals can be traced with the logic probe.

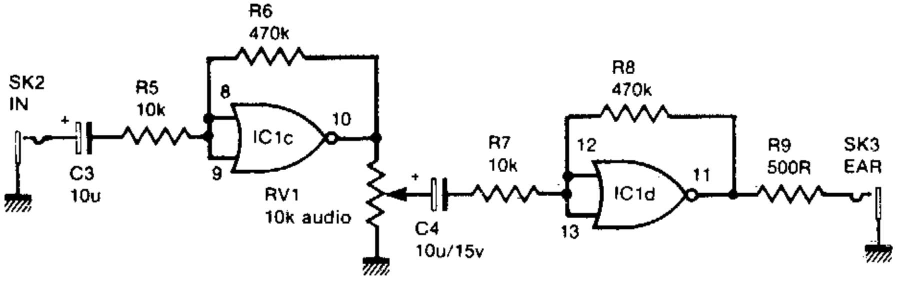
SIGNAL INJECTOR/TRACER



14
1
2
IC1a
3
7
-9V
R1
200k
R2
100k
5
6
IC1b
4
C2
0u1
C1
7n
R3
8k2
SK1
PROBE OUT
R4
1k
IC1 CD4001AE

Injector circuit diagram.

Fig. 67-2



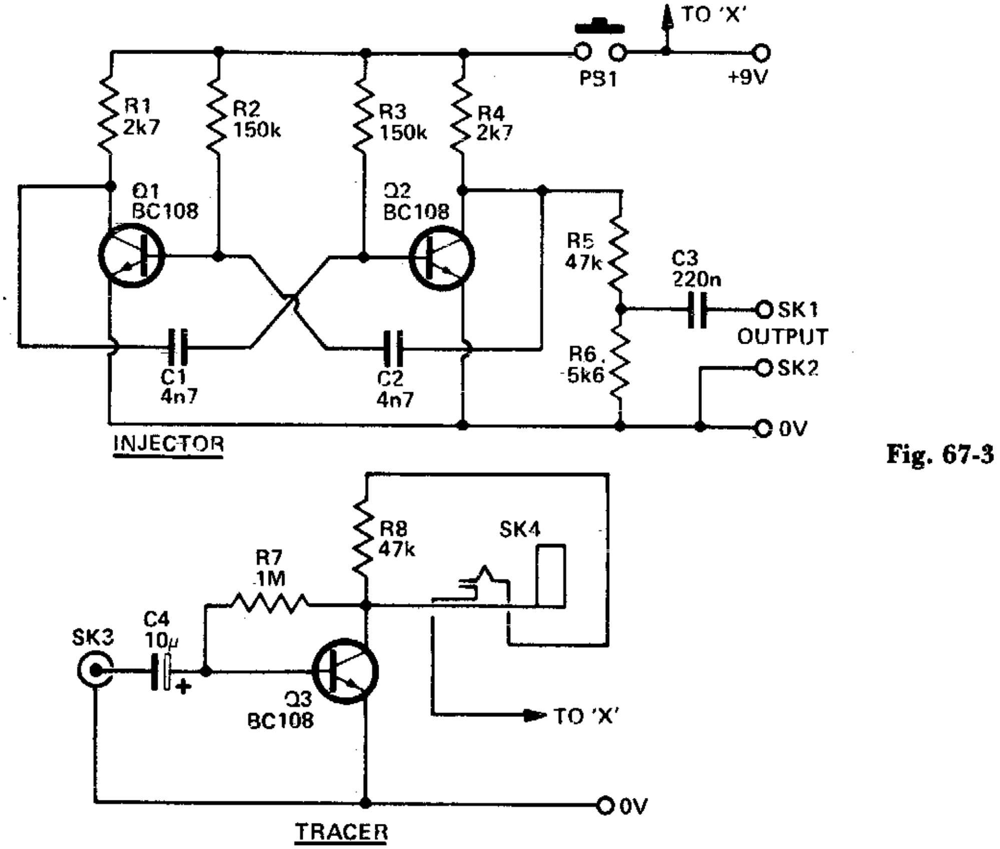
R6
470k
SK2
IN
+
C3
10u
R5
10k
8
9
IC1c
10
RV1
10k audio
+
C4
10u/15v
R7
10k
R8
470k
12
13
IC1d
11
R9
500R
SK3
EAR

Tracer circuit diagram.

Circuit Notes

The injector is a CMOS oscillator with period approximately equal to 1.4 × C1 × R2 seconds. The values are given for 1 kHz operation. Resistors R3 and R4 divide the output to 1 V. Whereas the oscillator employs the gates in their digital mode, the tracer used them in a linear fashion by applying negative feedback from output to input. They are used in much the same way as op amps. The circuit uses positive ground. It offers an advantage at the earphone output because one side of the earphone must be connected to ground via the case. Use of a positive ground allows the phone to be driven by the two N-channel transistors inside the CD4001 which are arranged in parallel and are thus able to handle more current for better volume.

INJECTOR/TRACER



TO 'X'
PB1
+9V
R1
2k7
R2
150k
R3
150k
R4
2k7
Q1
BC108
Q2
BC108
R5
47k
C3
220n
SK1
OUTPUT
SK2
R6
5k6
C1
4n7
C2
4n7
0V
INJECTOR
R8
47k
SK4
R7
1M
C4
10μ
SK3
Q3
BC108
TO 'X'
0V
TRACER

Fig. 67-3

The circuit diagrams for both parts of the injector/tracer. Note that SK4 is used to apply power to the amplifier section.

Circuit Notes

The unit has a separate amplifier and oscillator section allowing them to be used separately if need be. The injector is a multivibrator running at 1 kHz, with R5 and R6 dividing down the output to a suitable level (≈1 V). The tracer is a single-stage amplifier that drives the high impedance earpiece. C4 decouples the input.

CMOS LOGIC PROBE

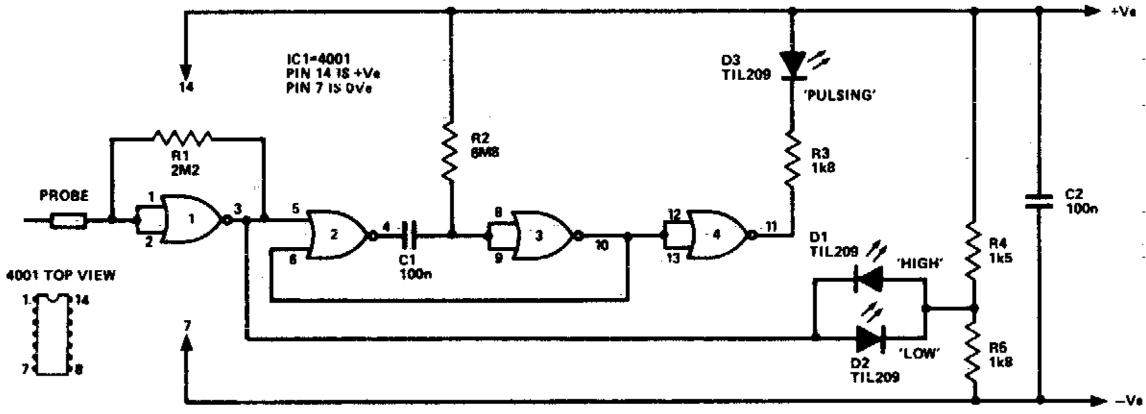

+Ve
IC1=4001
PIN 14 IS +Ve
PIN 7 IS 0Ve
14
D3
TIL209
'PULSING'
R1
2M2
R2
6M8
R3
1k8
PROBE
1
2
3
1
5
2
6
4
C1
100n
8
9
3
10
12
13
4
11
C2
100n
D1
TIL209
'HIGH'
R4
1k5
4001 TOP VIEW
1
14
7
8
7
'LOW'
D2
TIL209
R6
1k8
–Ve

Fig. 67-4

Circuit Notes

The logic probe can indicate four input states, as follows: floating input—all LEDs off; logic 0 input—D2 switched on (D3 will briefly flash on); logic 1 input—D1 switched on; pulsing input—D3 switched on, or pulsing in the case of a low frequency input signal (one or both of the other indicators will switch on, showing if one input state predominates).

RF PROBE FOR VOM

PARTS LIST FOR
RF PROBE FOR VOM

C1—500-pF, 400-VDC capacitor
C2—0.001-uF, disc capacitor
D1—1N4149 diode
R1—15,000-ohm, ½-watt resistor

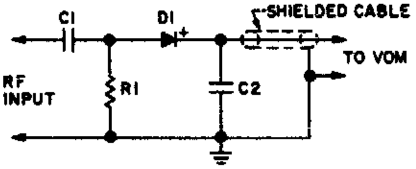

CI
DI
+
SHIELDED CABLE
TO VOM
RF
INPUT
RI
C2

Fig. 67-5

Circuit Notes

This probe makes possible relative measurements of rf voltages to 200 MHz on a 20,000 ohms-per-volt multimeter. Rf voltage must not exceed the breakdown rating of the 1N4149—approximately 100 V.

100 K MEGOHM DC PROBE

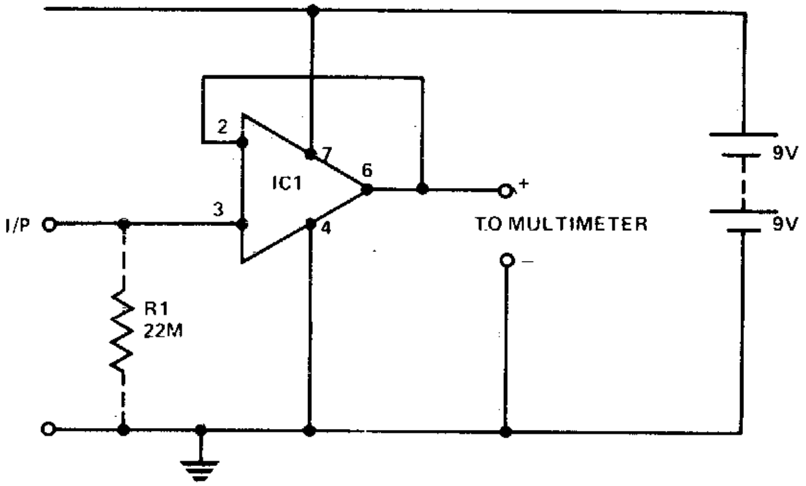

2
7
IC1
6
+
3
4
I/P
TO MULTIMETER
−
9V
9V
R1
22M

Fig. 67-6

Circuit Notes

A 741 op amp is used with 100% ac and dc feedback to provide a typical input impedance of 10^{11} ohm and unity gain. To avoid hum and rf pickup the input leads should be kept as short as possible and the circuit should be mounted in a small grounded case. Output leads may be long since the output impedance of the circuit is a fraction of an ohm. With no input the output level is indeterminate. Including R1 in the circuit through lowers the input impedance to 22 M.

AUDIBLE TTL PROBE

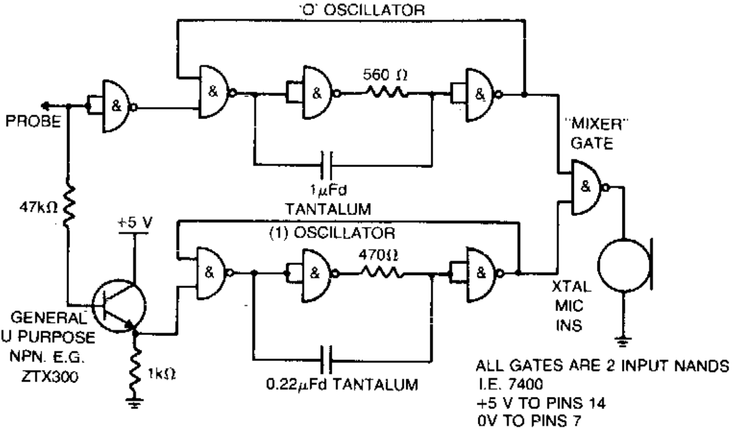

'O' OSCILLATOR
560 Ω
&
PROBE
"MIXER"
GATE
1μFd
TANTALUM
47kΩ
+5 V
(1) OSCILLATOR
470Ω
XTAL
MIC
INS
GENERAL PU PURPOSE
NPN. E.G.
ZTX300
1kΩ
0.22μFd TANTALUM
ALL GATES ARE 2 INPUT NANDS
I.E. 7400
+5 V TO PINS 14
0V TO PINS 7

Fig. 67-7

Circuit Notes

When the probe is in contact with a TTL low (0) the probe emits a low note. With a TTL high (1), a high note is emitted. Power is supplied by the circuit under test.

LOGIC PROBE

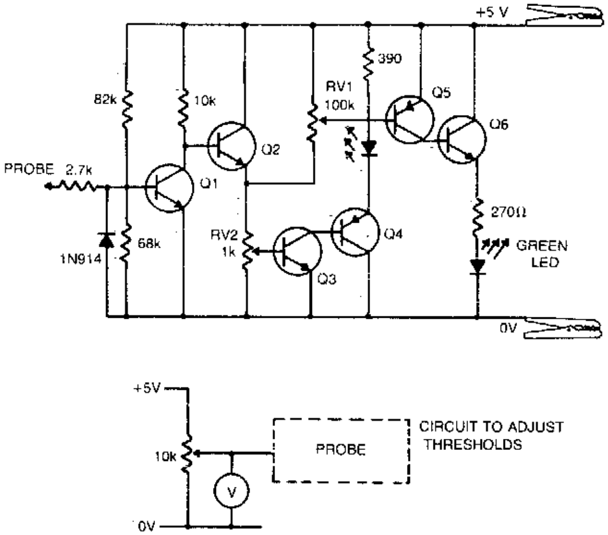

+5 V
390
RV1
100k
Q5
Q6
82k
10k
Q2
PROBE 2.7k
Q1
270Ω
GREEN
LED
1N914
68k
RV2
1k
Q4
Q3
0V
+5V
10k
PROBE
CIRCUIT TO ADJUST
THRESHOLDS
V
0V

Fig. 67-8

Circuit Notes

Transistors Q1 and Q2 form a buffer, providing the probe with a reasonable input impedance. Q3 and Q4 form a level detecting circuit. As the voltage across the base-emitter junction of the Q3 rises above 0.6 V the transistor turns on thus turning on Q4 and lighting the red (high) LED. Q5 and Q6 perform the same function but for the green (low) LED. Q1, Q4, Q5 are all PNP general purpose silicon transistors (BC178 etc). Q2, Q3, Q6 are all PNP general purpose silicon transistors (BC 108 etc.) The threshold low is ≤ 0.8 V, and the threshold high is ≥ 2.4 V.

LOGIC TEST PROBE WITH MEMORY

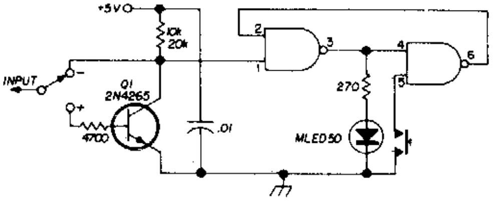

+5V
10k
20k
2
3
4
6
1
5
INPUT
Q1
2N4265
270
.01
4700
MLED50

Fig. 67-9

Circuit Notes

There are two switches: a memory disable switch and a pulse polarity switch. Memory disable is a push-button that resets the memory to the low state when depressed. Pulse polarity is a toggle switch that selects whether the probe responds to a high-level or pulse (+5 V) or a low-level or pulse (ground). (Use IC logic of the same type as is being tested).

LOGIC PROBE

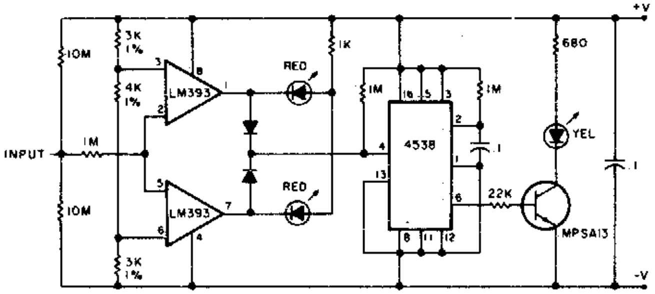

+V
10M
3K 1%
4K 1%
3
8
2
LM393
1
RED
1K
INPUT
1M
5
6
LM393
7
4
10M
3K 1%
RED
1M
16
5
3
1M
2
4538
4
.1
1
13
6
8
11
12
680
YEL
22K
MPSA13
.1
-V

Fig. 67-10

Circuit Notes

The probe indicates a high or low at 70% and 30% of V+ (5 to 12 V). One section of the voltage comparator (LM393) senses V in over 70% of supply and the second section senses V in under 30%. These two sections direct-drive the appropriate LEDs. The pulse detector is a CMOS oneshot (MC14538) triggered on the rising edge of the LM393 outputs through 1N4148 diodes. With the RC values shown, it triggered reliably at greater than 30 kHz on both sine and square waves.

SIMPLE LOGIC PROBE

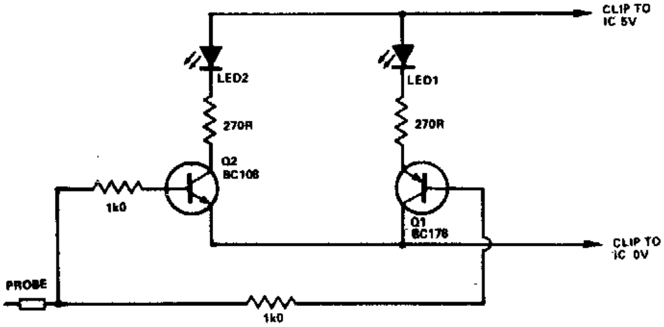
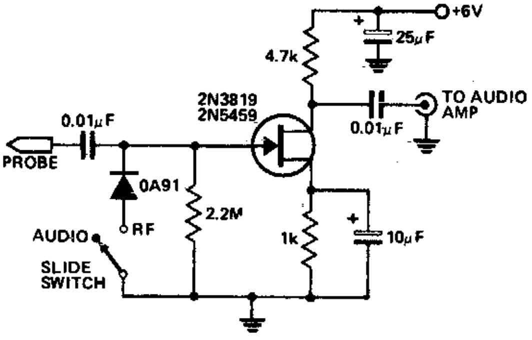

CLIP TO IC 5V
LED2
270R
Q2 BC108
1k0
LED1
270R
Q1 BC178
CLIP TO IC 0V
PROBE
1k0

Fig. 67-11

Circuit Notes

If the probe is connected to logic 0, Q1 will be turned on lighting D1. At logic 1, Q2 will be turned on lighting D2. For Q1 and Q2 any NPN or PNP transistors will do. Similarly, D1 and D2 can be any LEDs.

AUDIO-RF SIGNAL TRACER PROBE


+6V
25μF
4.7k
2N3819
2N5459
0.01μF
TO AUDIO AMP
0.01μF
PROBE
0A91
2.2M
RF
AUDIO
SLIDE SWITCH
1k
10μF

Circuit Notes

This economical signal tracer is useful for servicing and alignment work in receivers and low power transmitters. When switched to RF, the modulation on any signal is detected by the diode and amplified by the FET. A twin-core shielded lead can be used to connect the probe to an amplifier and to feed 6 volts to it.

Fig. 67-12

TTL LOGIC TESTER

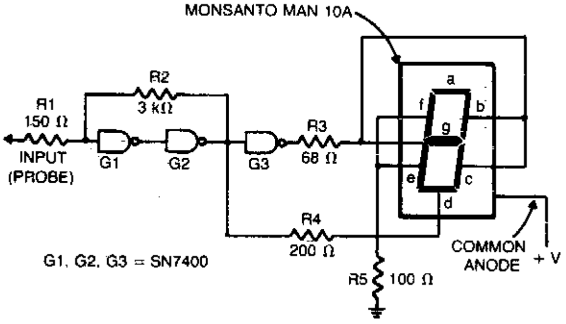

MONSANTO MAN 10A
R2
3 kΩ
R1
150 Ω
INPUT
(PROBE)
G1
G2
G3
R3
68 Ω
a
b
c
d
e
f
g
R4
200 Ω
R5
100 Ω
COMMON ANODE
+V
G1, G2, G3 = SN7400

Fig. 67-13

Circuit Notes

Gates G1 and G2 together with resistors R1 and R2 form a simple voltage monitor that has a trip point of 1.4 volts. Gate G3 is simply an inverter. The display section of the tester consists of a common anode alphanumeric LED and current-limiting resistors. It indicates whether the input voltage is above or below 1.4 V, and displays a H or a L (for high or low logic-level) respectively.

68

Pulse Generators

The sources of the following circuits are contained in the Sources section beginning on page 730. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

Pulse Generator
Single Op Amp Oscillator
Programmable Pulse Generator
Unijunction Transistor Pulse Generators
Pulse Generator

Pulse Generator
Free-Running Oscillator
Pulse Generator with 25% Duty Cycle
Pulse Generator
555 Timer Oscillator

Versatile Two-Phase Pulse Generator

PULSE GENERATOR

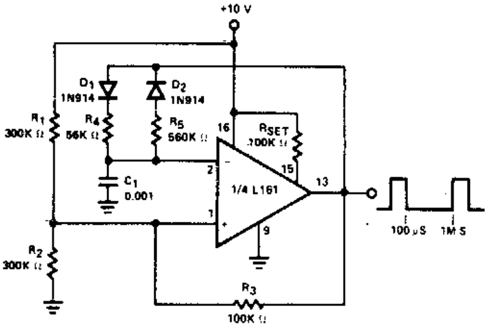

+10 V
D1
1N914
D2
1N914
R1
300K Ω
R4
56K Ω
R5
560K Ω
16
RSET
100K Ω
2
15
13
1/4 L161
C1
0.001
1
9
100 μS
1M S
R2
300K Ω
R3
100K Ω

Fig. 68-1

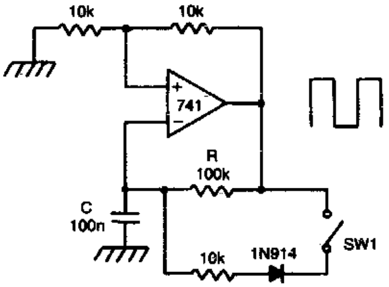
Circuit Notes

The duty cycle of the output pulse is equal to R4/(R4 + R5) × 100%. For duty cycles of less than 50%, D1 can be eliminated and R2 raised according to the following formula:

$$R4(\text{actual}) = \frac{R5 \times R4(\text{eff})}{R5 - R4(\text{eff})}$$

R4(eff) is the effective value of R4 in the circuit and R4(actual) is the actual value used; R4(actual) will always be larger than R4(eff).

SINGLE OP AMP OSCILLATOR

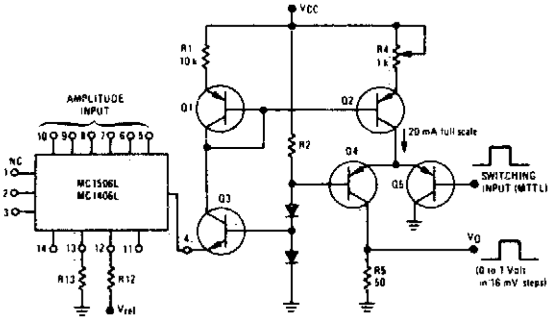

10k
10k
+
741
−
R
100k
C
100n
10k
1N914
SW1

Circuit Notes

This circuit has a Schmitt trigger and integrator built around one op amp. Timing is controlled by the RC network. Voltage at the inverting input follows the RC charging exponential within the upper and lower hysteresis levels. By closing the switch SW1, the discharge time of the capacitor becomes ten times as fast as the rise time. Thus a square wave with an 10:1 mark space ratio is generated.

Fig. 68-2

PROGRAMMABLE PULSE GENERATOR

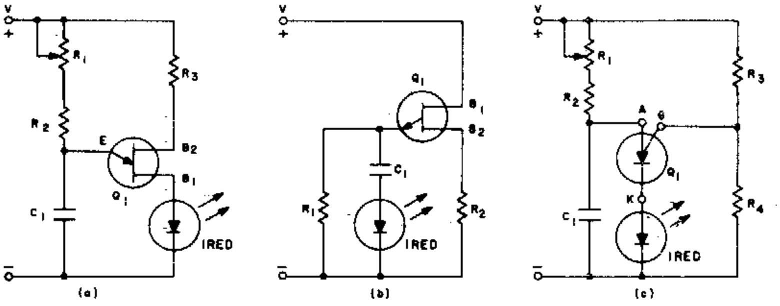

VCC
R1
10k
R4
1k
AMPLITUDE
INPUT
Q1
Q2
20 mA full scale
10
9
8
7
6
5
R2
Q4
NC
1
2
3
MC1506L
MC1406L
Q3
Q5
SWITCHING
INPUT (MTTL)
4
14
13
12
11
Vo
R13
R12
Vref
R5
50
(0 to 1 Volt
in 16 mV steps)

Circuit Notes

Fast rise and fall times require the use of high speed switching transistors for the differential pair, Q4 and Q5. Linear ramps and sine waves may be generated by the appropriate reference input.

Fig. 68-3

UNIJUNCTION TRANSISTOR PULSE GENERATORS


V
+
R1
R3
R2
E
B2
B1
Q1
C1
IRED
(a)
V
+
Q1
B1
B2
C1
R1
R2
IRED
(b)
V
+
R1
R3
R2
A
G
Q1
K
C1
R4
IRED
(c)

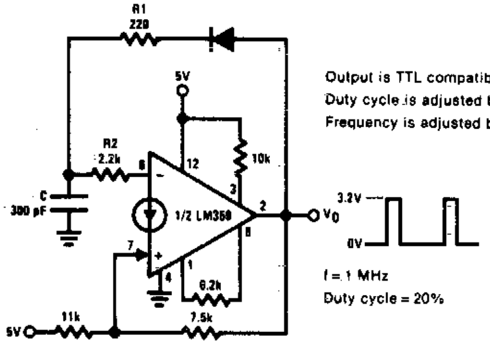
(a) Pulser With Unijunction Transistor

(b) Pulser With Complementary Unijunction Transistor

(c) Pulser With Programmable Unijunction Transistor

Fig. 68-4

PULSE GENERATOR

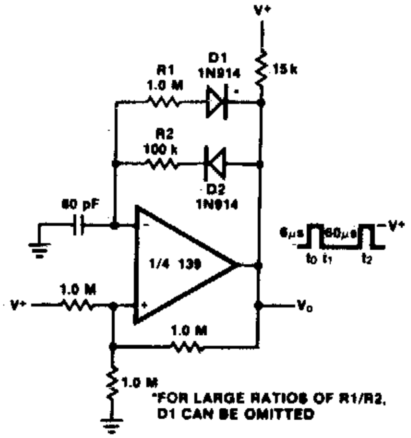

R1
220
5V
R2
2.2k
6
12
10k
3
2
C
300 pF
1/2 LM358
8
7
4
1
6.2k
5V
11k
7.5k
V0
3.2V
0V

Output is TTL compatible
Duty cycle is adjusted by R1
Frequency is adjusted by C

f = 1 MHz
Duty cycle = 20%

Fig. 68-5

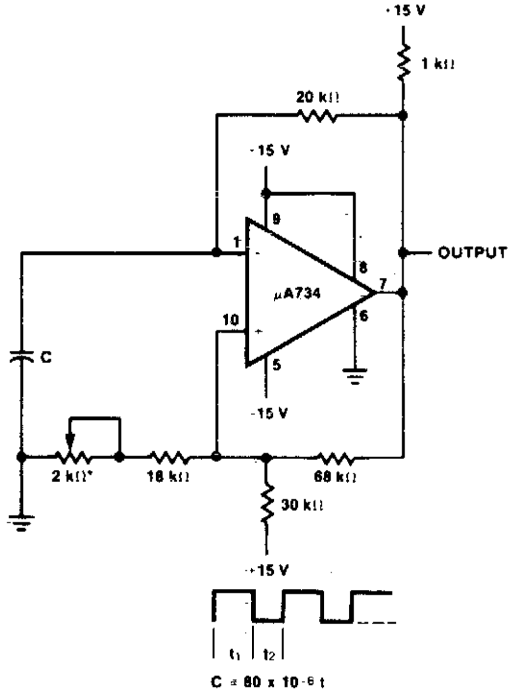
PULSE GENERATOR


V+
D1
R1
1.0 M
1N914
15 k
R2
100 k
D2
1N914
60 pF
1/4 139
6µs
60µs
V+
t0
t1
t2
V+
1.0 M
V0
1.0 M
1.0 M

*FOR LARGE RATIOS OF R1/R2, D1 CAN BE OMITTED

Fig. 68-6

FREE-RUNNING OSCILLATOR


+15 V
1 kΩ
20 kΩ
-15 V
9
1
-
μA734
8
7
OUTPUT
10
+
6
5
-15 V
C
2 kΩ*
18 kΩ
68 kΩ
30 kΩ
+15 V
t1
t2
C = 80 x 10-6 t

Adjusts $\frac{T_1}{T_2}$

Fig. 68-7

PULSE GENERATOR

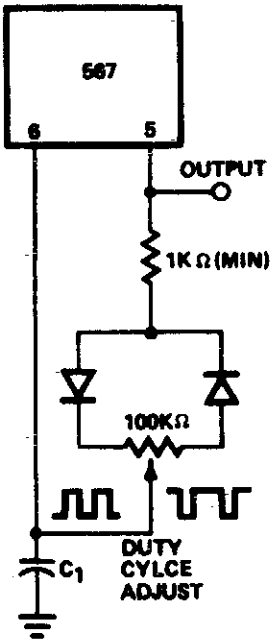

567
6
5
OUTPUT
1K Ω (MIN)
100KΩ
C1
DUTY
CYLCE
ADJUST

Fig. 68-9

PULSE GENERATOR WITH 25% DUTY CYCLE

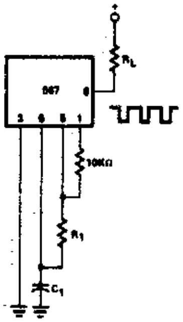
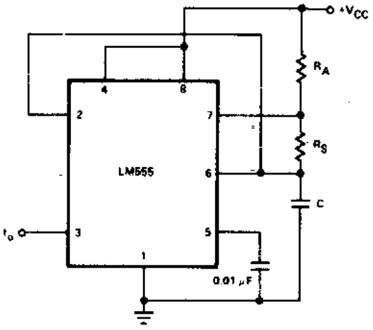

+
RL
567
8
3
6
5
1
10KΩ
R1
C1

Fig. 68-8

555 TIMER OSCILLATOR

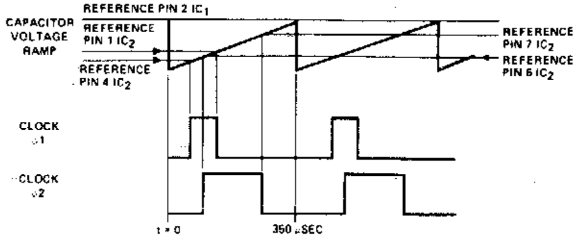

+VCC
RA
RS
C
4
8
2
7
LM555
6
fo
3
5
1
0.01 μF

$$f = \frac{1.44}{(R_A + 2\,R_B)\,C}\,, \quad \text{duty cycle} = \frac{R_B}{R_A + 2\,R_B}$$

a. f = 120 kHz, C = 1200 pF, $R_A = R_B = 10\ k\,\Omega$

Fig. 68-10

VERSATILE TWO-PHASE PULSE GENERATOR



Typical Waveforms of the Two Phase Pulse Generator

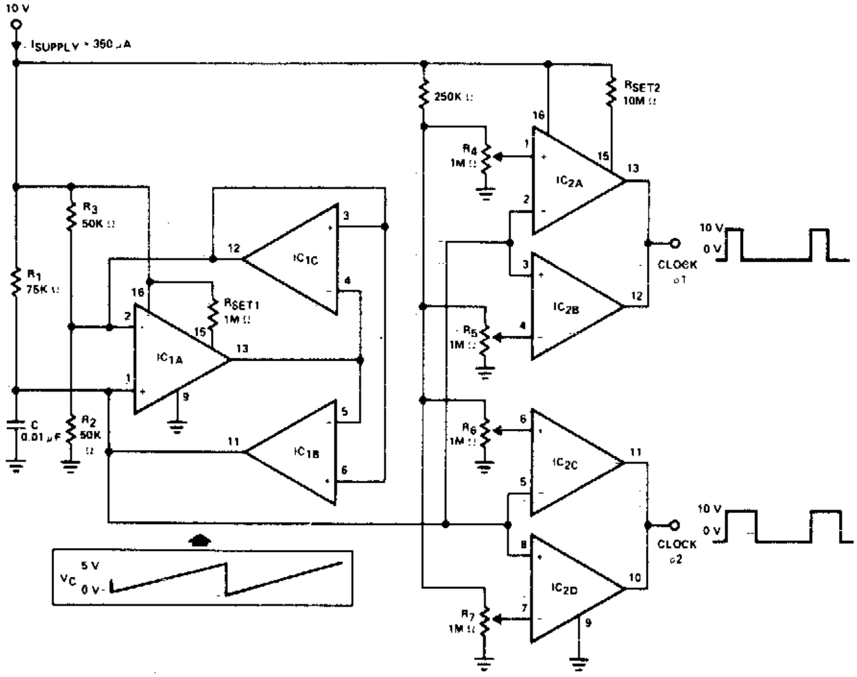

10 V
I SUPPLY = 350 μA
250K Ω
RSET2 10M Ω
R4 1M Ω
R3 50K Ω
R1 75K Ω
RSET1 1M Ω
IC1A
IC1B
IC1C
IC2A
IC2B
IC2C
IC2D
C 0.01 μF
R2 50K Ω
R5 1M Ω
R6 1M Ω
R7 1M Ω
CLOCK φ1
CLOCK φ2
10 V
0 V
VC
5 V
0 V

Fig. 68-11

Circuit Notes

Two-phase clock generator uses two L161s to generate pulses of adjustable widths and phase relationships. Ramp generator feeds two variable window comparators formed by IC2A-IC2B and IC2C-IC2D respectively.

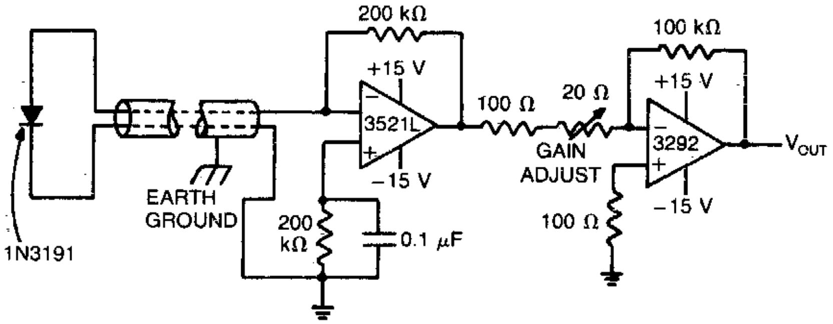
69

Radiation Detectors

The sources of the following circuits are contained in the Sources section beginning on page 730. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

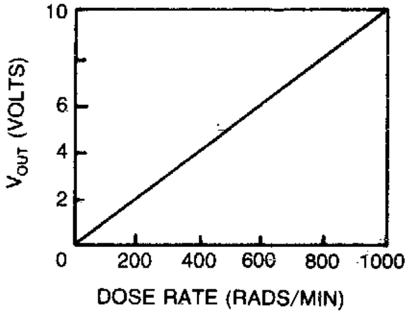
Dosage-Rate Meter
Wideband Radiation Monitor
Gamma Ray Pulse Integrator
Sensitive Geiger Counter
Geiger Counter
Nuclear Particle Detector

DOSAGE-RATE METER



200 kΩ
100 kΩ
+15 V
+15 V
3521L
100 Ω
20 Ω
GAIN ADJUST
3292
V_{OUT}
−15 V
−15 V
EARTH GROUND
200 kΩ
0.1 μF
100 Ω
1N3191

Fig. 69-1



Circuit Notes

A commercial diode is the detector in this highly accurate radiation monitor. The lowdrift FET-input op amp amplifies detector current to a usable level, and the chopper-stabilized amplifier then provides additional gain while minimizing any error caused by ambient-temperature fluctuations. Gain is adjusted so that the output voltage is 1% of incident radiation intensity in rads per minute; therefore voltage can be displayed on 3½ digit DVM for direct reading of dosage rate. Output voltage from the monitor is linearly proportional to radiation intensity at the diode.

WIDEBAND RADIATION MONITOR

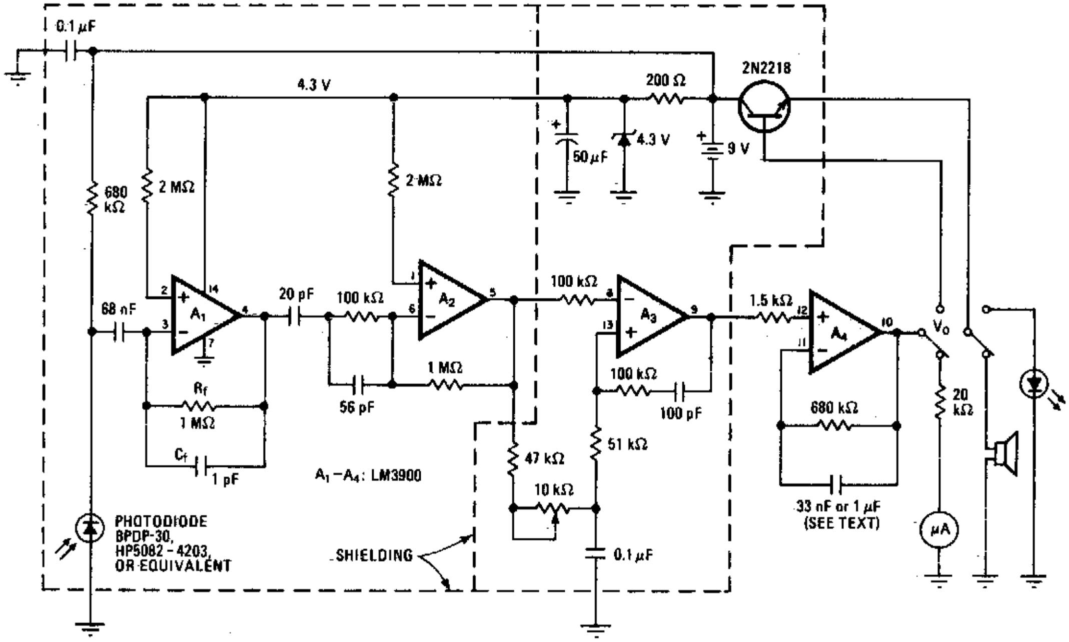
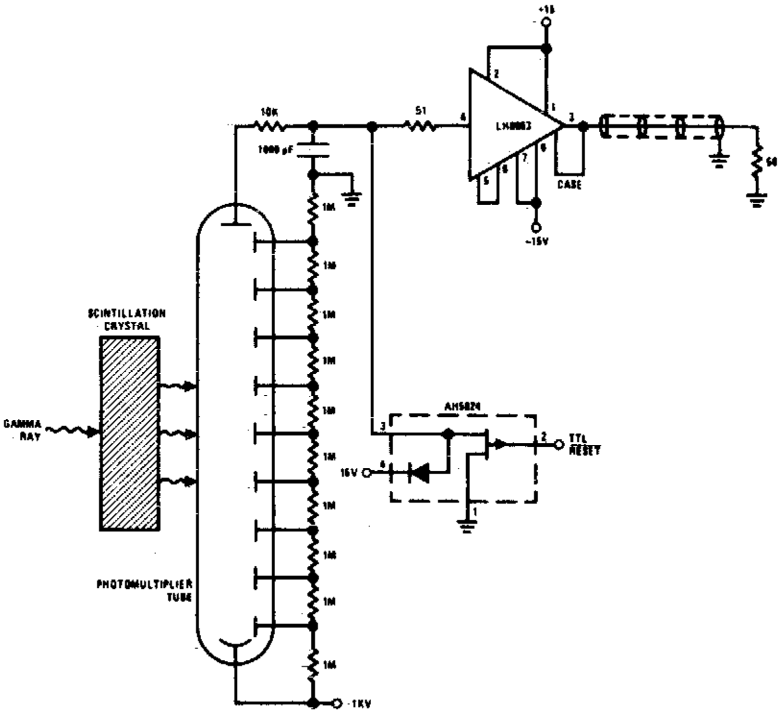

0.1 μF
4.3 V
200 Ω
2N2218
50 μF
4.3 V
9 V
680 kΩ
2 MΩ
2 MΩ
68 nF
A₁
20 pF
100 kΩ
A₂
100 kΩ
A₃
1.5 kΩ
A₄
V₀
R_f
1 MΩ
56 pF
1 MΩ
100 kΩ
100 pF
680 kΩ
20 kΩ
C_f
1 pF
A₁ – A₄: LM3900
47 kΩ
51 kΩ
10 kΩ
33 nF or 1 μF (SEE TEXT)
μA
PHOTODIODE BPDP-30, HP5082 – 4203, OR EQUIVALENT
SHIELDING
0.1 μF

Fig. 69-2

Circuit Notes

A sensitive radiation monitor may be simply constructed with a large-area photodiode and a quad operational amplifier. Replacing the glass window of the diode with Mylar foil will shield it from light and infrared energy, enabling it to respond to such nuclear radiation as alpha and beta particles and gamma rays. A4 integrates the output of A3 in order to drive a microammeter. A 1 microfarad capacitor is used in the integrating network. A lower value, say, 33 nanofarads, will make it possible to drive a small loudspeaker (50-hertz output signal) or light-emitting diode.

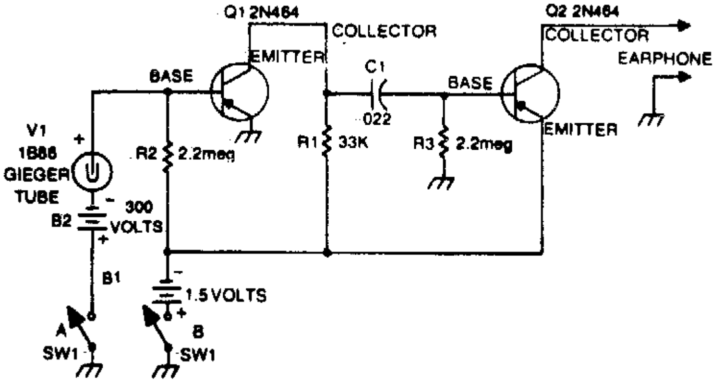
GAMMA RAY PULSE INTEGRATOR



+15
2
1
10K
51
4
LH0063
3
1000 pF
5
6
7
8
CASE
50
-15V
1M
1M
1M
1M
1M
1M
1M
1M
1M
1M
SCINTILLATION CRYSTAL
GAMMA RAY
AH5024
3
2
TTL RESET
15V
4
1
PHOTOMULTIPLIER TUBE
-1KV

Fig. 69-3

SENSITIVE GEIGER COUNTER



Q1 2N464
COLLECTOR
EMITTER
BASE
C1
.022
BASE
Q2 2N464
COLLECTOR
EARPHONE
EMITTER
V1
1B86
GIEGER
TUBE
+
-
R2 2.2meg
R1 33K
R3 2.2meg
B2
300
VOLTS
+
B1
-
1.5 VOLTS
+
A
SW1
B
SW1

Fig. 69-4

GEIGER COUNTER

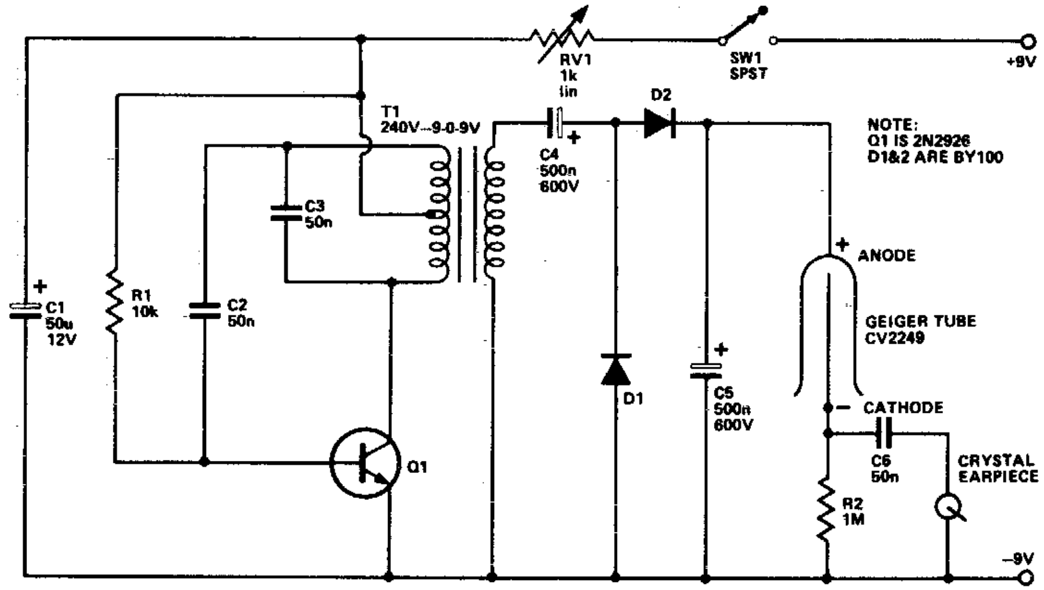

RV1
1k
lin
SW1
SPST
+9V
T1
240V–9-0-9V
C4
500n
600V
D2
NOTE:
Q1 IS 2N2926
D1&2 ARE BY100
+
ANODE
C3
50n
+
C1
50u
12V
R1
10k
C2
50n
GEIGER TUBE
CV2249
D1
+
C5
500n
600V
– CATHODE
C6
50n
CRYSTAL
EARPIECE
Q1
R2
1M
–9V

Fig. 69-5

Circuit Notes

The Geiger tube needs a high voltage supply which consists of Q1 and its associated components. The transformer is connected in reverse; the secondary is connected as a Hartley oscillator, and R1 provides base bias. D1, D2, C4, and C5 comprise a voltage doubler. RV1 should be set so that each click heard is nice and clean because over a certain voltage range all that will be heard is a continuous buzz.

NUCLEAR PARTICLE DETECTOR

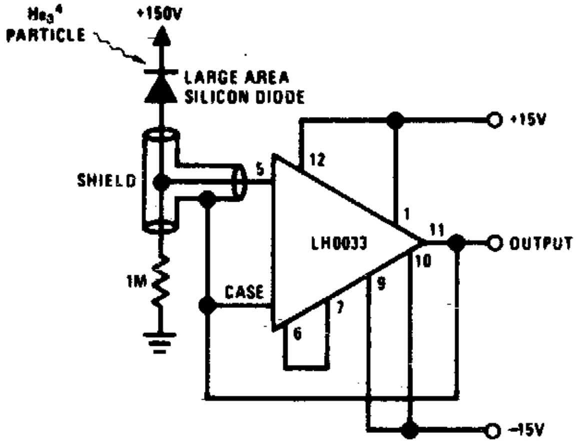

He_3^4
PARTICLE
+150V
LARGE AREA
SILICON DIODE
+15V
SHIELD
5
12
1
11
LH0033
10
OUTPUT
1M
CASE
9
7
6
–15V

Fig. 69-6

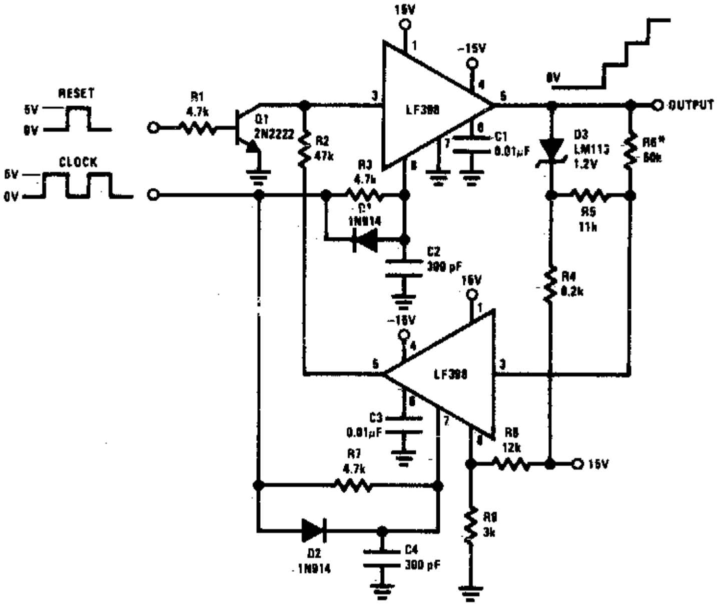
70

Ramp Generators

The sources of the following circuits are contained in the Sources section beginning on page 730. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

Staircase Generator
Linear Voltage Ramp Generator
Precision Ramp Generator
Ramp Generator with Variable Reset Level

STAIRCASE GENERATOR


15V
-15V
RESET
5V
0V
CLOCK
5V
0V
R1
4.7k
Q1
2N2222
R2
47k
R3
4.7k
D1
1N914
C2
300 pF
LF398
1
3
4
5
6
7
8
C1
0.01µF
0V
OUTPUT
D3
LM113
1.2V
R6*
50k
R5
11k
R4
8.2k
15V
-15V
LF398
C3
0.01µF
R7
4.7k
R8
12k
15V
R9
3k
D2
1N914
C4
300 pF

*Select for step height
50k → ≈ 1V Step

Fig. 70-1

LINEAR VOLTAGE RAMP GENERATOR

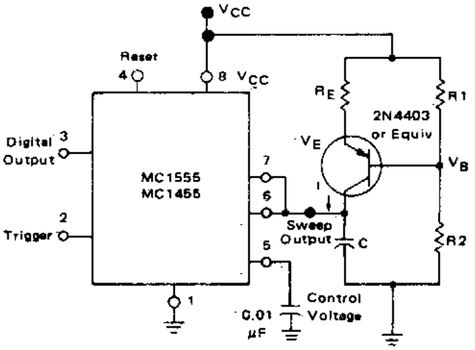

Vcc
Reset
4
8 Vcc
Digital Output 3
2
Trigger
MC1555
MC1455
7
6
5
1
RE
R1
2N4403 or Equiv
VE
VB
I
Sweep Output
C
R2
Control Voltage
0.01 µF

Fig. 70-2

Circuit Notes

In the monostable mode, the resistor can be replaced by a constant current source to provide a linear ramp voltage. The capacitor still charges from 0 to 2/3 Vcc. The linear ramp time is given by the following equation:

$$I = \frac{V_{CC} - V_B - V_{BE}}{R_E} \qquad t = \frac{2}{3}\frac{V_{CC}}{I}$$

If V_B is much larger than V_{BE}, then t can be made independent of V_{CC}.

PRECISION RAMP GENERATOR

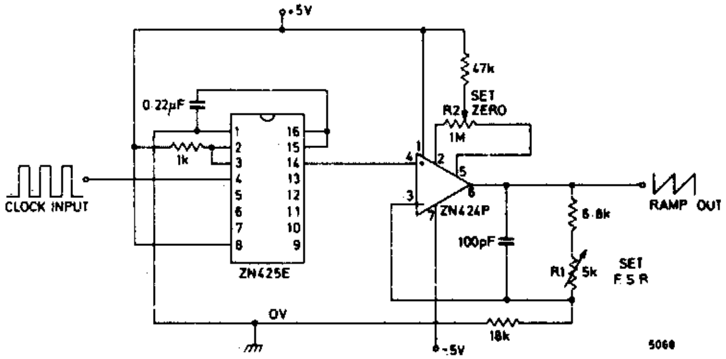
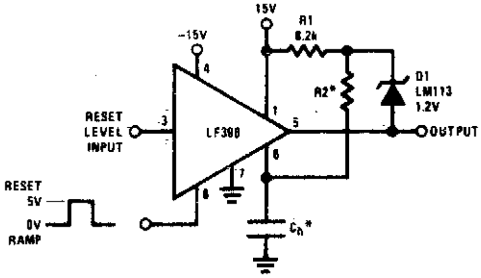

+5V
47k
SET
ZERO
R2
1M
0.22μF
1k
CLOCK INPUT
1
2
3
4
5
6
7
8
16
15
14
13
12
11
10
9
ZN425E
4
1
2
5
3
6
7
ZN424P
RAMP OUT
6.8k
100pF
R1
5k
SET
F.S.R
0V
18k
-5V
5060

Fig. 70-3

RAMP GENERATOR WITH VARIABLE RESET LEVEL


15V
R1
8.2k
-15V
4
D1
LM113
1.2V
R2*
RESET
LEVEL
INPUT
3
LF398
1
5
OUTPUT
6
7
RESET
5V
0V
RAMP
8
Ch*

*Select for ramp rate
R ≥ 10k

$$\frac{\Delta V}{\Delta T} = \frac{1.2V}{(R2)(C_h)}$$

Fig. 70-4

71

Receivers

The sources of the following circuits are contained in the Sources section beginning on page 730. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

- Clock Radio
- AM/FM Clock Radio
- AM Radio
- FM Stereo Demodulation System
- Analog Receiver
- FM Radio
- Simple LF Converter
- CMOS Line Receiver
- Squelch Circuit for AM or FM
- VLF Converter

CLOCK RADIO

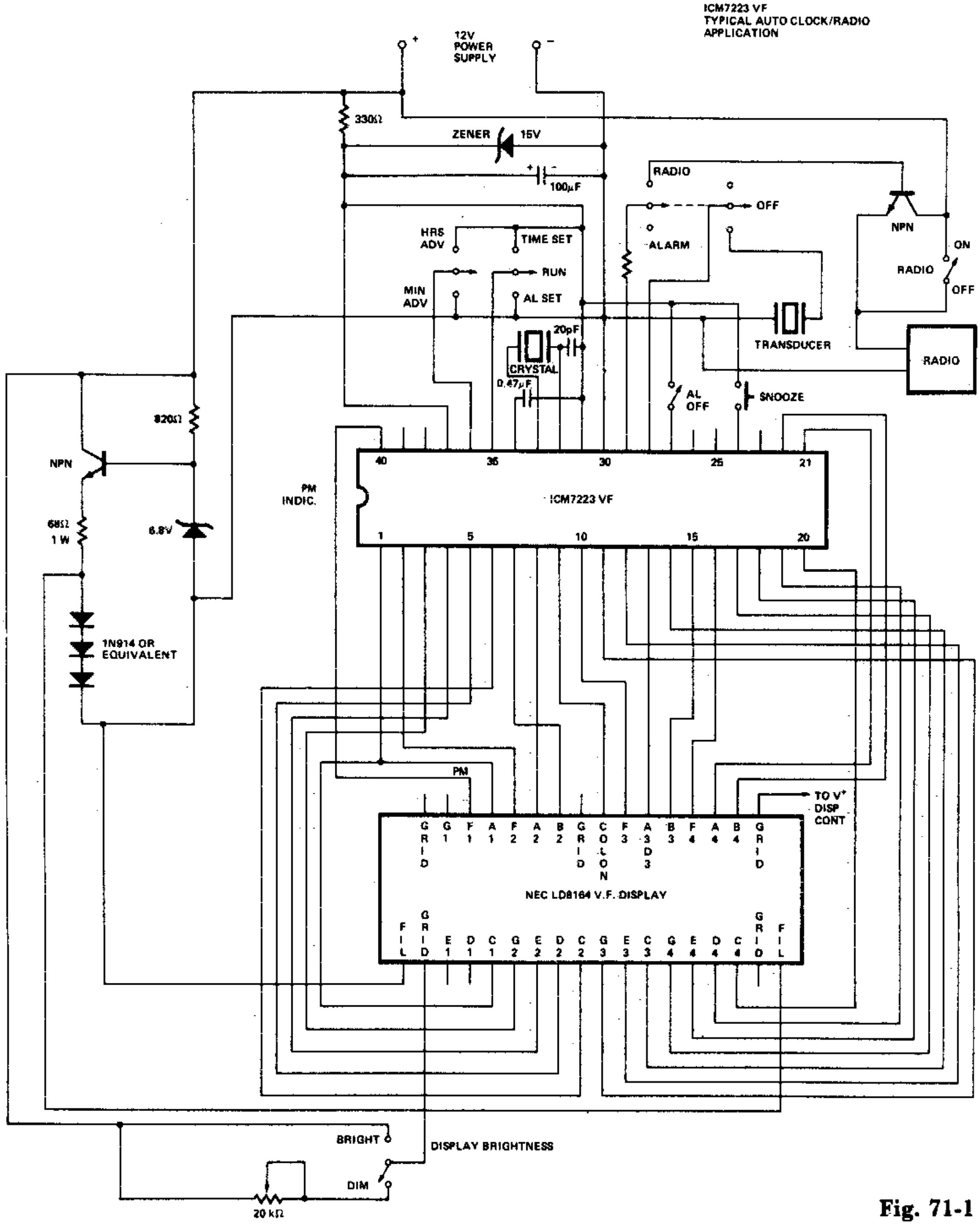
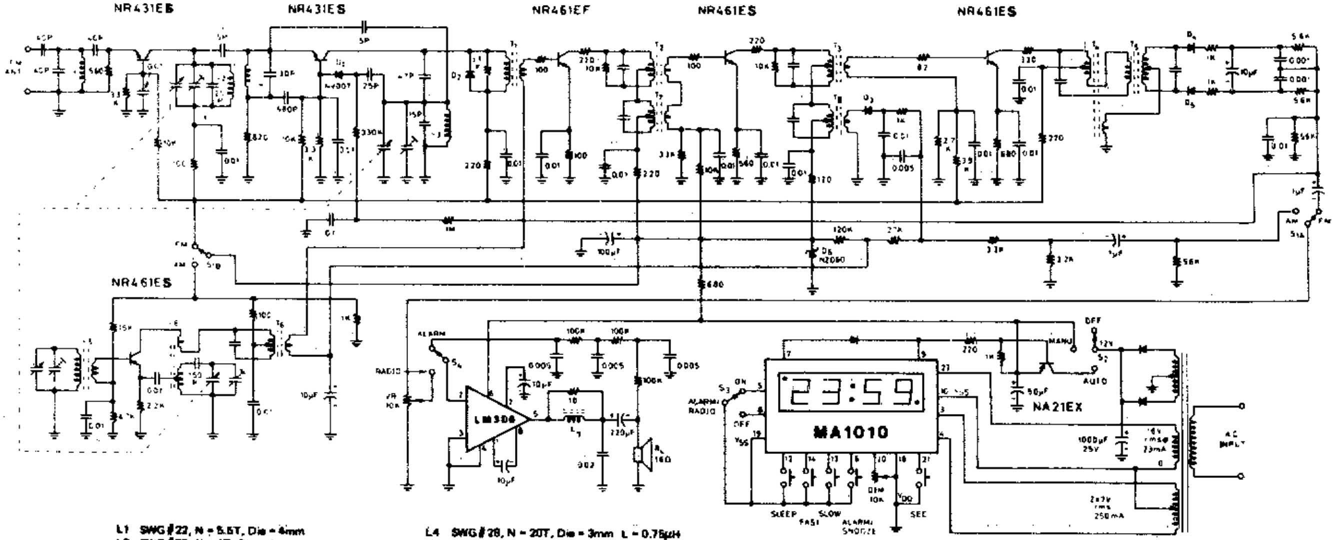

ICM7223 VF
TYPICAL AUTO CLOCK/RADIO
APPLICATION
+
12V
POWER
SUPPLY
−
330Ω
ZENER
15V
+
−
100μF
RADIO
OFF
ALARM
NPN
ON
RADIO
OFF
HRS
ADV
TIME SET
MIN
ADV
RUN
AL SET
20pF
CRYSTAL
0.47μF
TRANSDUCER
RADIO
AL
OFF
SNOOZE
820Ω
NPN
68Ω
1 W
6.8V
1N914 OR
EQUIVALENT
PM
INDIC.
40
35
30
25
21
ICM7223 VF
1
5
10
15
20
PM
TO V+
DISP
CONT
GRID
G1
F1
A1
F2
A2
B2
GRID
COLON
F3
A3
D3
B3
F4
A4
B4
GRID
NEC LD8164 V.F. DISPLAY
FIL
GRID
E1
D1
C1
G2
E2
D2
C2
G3
E3
C3
G4
E4
D4
C4
GRID
FIL
BRIGHT
DISPLAY BRIGHTNESS
DIM
20 kΩ

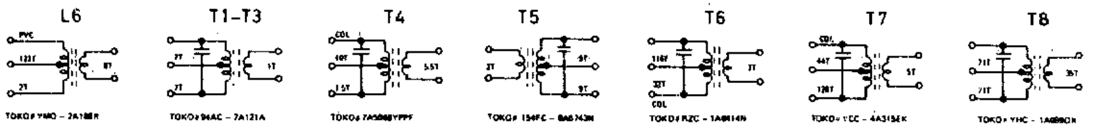
Fig. 71-1

AM/FM CLOCK RADIO


NR431EB
NR431ES
NR461EF
NR461ES
NR461ES
NR461ES
FM ANT
D2
D3
D4
D5
D6
NZ060
T1
T2
T3
T4
T5
T6
T7
T8
L1
L2
L3
L5
L6
L7
S1A
S1B
S2
S3
S4
FM
AM
ALARM
RADIO
VR 10k
LM386
MA1010
23:59
NA21EX
MANU
AUTO
OFF
12V
ON
ALARM/RADIO
SLEEP
FAST
SLOW
ALARM/SNOOZE
SEC
DIM 10k
VDD
VSS
1000μF 25V
AC INPUT
16Ω
220μF
10μF
100μF
1μF

L1 SWG #22, N = 5.5T, Dia = 4mm
L2 SWG #22, N = 4T, Dia = 4mm
L3 SWG #22, N = 4T, Dia = 3mm

L4 SWG #28, N = 20T, Dia = 3mm L = 0.75μH
L5 95T : 8T, L = 600μH, Qμ = 300
L7 N = 1.5T, PHILIPS #4312-020-34401


L6
T1–T3
T4
T5
T6
T7
T8
PVC
121T
2T
8T
7T
1T
COL
40T
1.5T
5.5T
9T
116T
32T
3T
44T
120T
71T
35T
TOKO# YMO – 2A18ER
TOKO# 94AC – 7A121A
TOKO# 7A5068YPPF
TOKO# 154FC – 8A6743N
TOKO# RZC – 1A8814N
TOKO# YCC – 4A3152K
TOKO# YHC – 1A8880K

FM performance (88–108 MHz)

30dB quieting sensitivity:	5μV
limiting sensitivity:	20μV
AM rejection:	40dB
AFC holding range:	800KHz
Bandwidth:	180KHz

AM performance (525–1650 KHz)

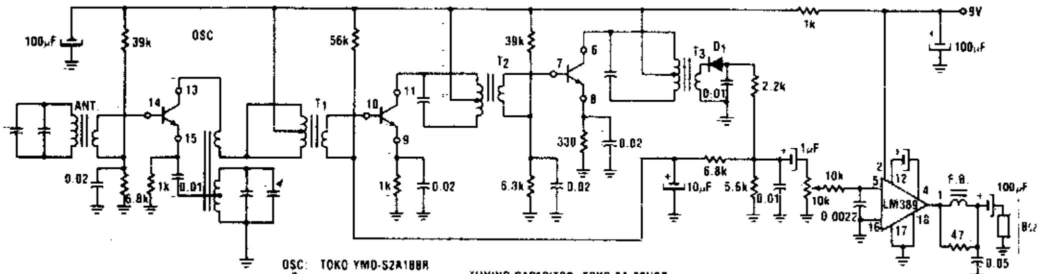
maximum sensitivity:	100μV/M
20dB quieting sensitivity:	280μV/M
selectivity ± 10KHz:	–28dB
AGC figure of merit:	40dB
overload distortion:	6%

AUDIO performance

gain at 1 KHz:	200
10% THD output power:	900mW
frequency response:	70Hz – 12KHz
typical system dist:	0.8%
alarm tone frequency:	600Hz

Fig. 71-2

AM RADIO



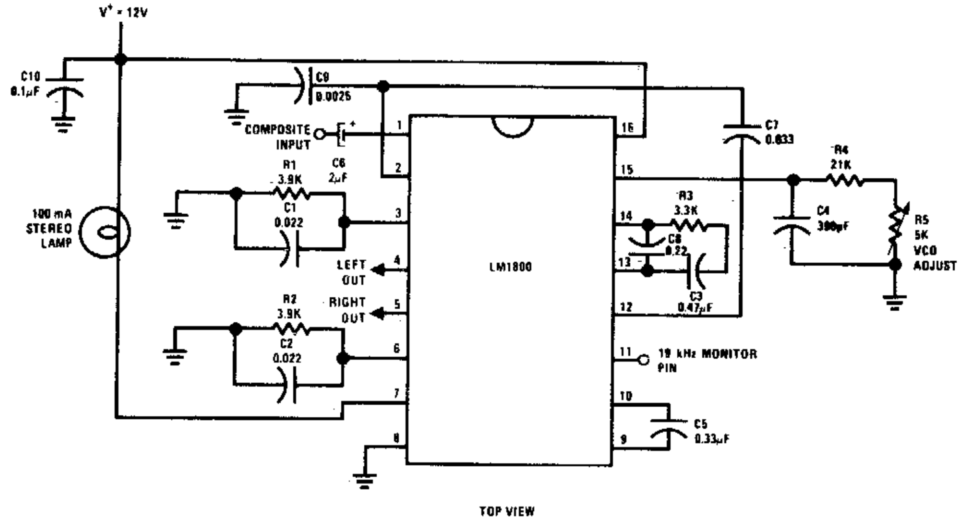
100μF
39k
OSC
56k
39k
T2
T3
D1
1k
9V
100μF
ANT.
14
13
15
T1
10
11
9
7
6
8
2.2k
0.01
330
0.02
6.8k
1μF
0.02
6.8k
1k
0.01
1k
0.02
6.3k
0.02
10μF
5.6k
0.01
10k
10k
0.0022
2
12
5
LM389
4
18
16
17
F.B.
1
100μF
47
0.05
8Ω

OSC: TOKO YMD-52A188R
T1:
T2:
T3: TOKO RHC-1A5006DX
ANT: L 650μH, Qu 250

*TUNING CAPACITOR: TOKO 2A-20HQZ
FERRITE BEAD. PHILIPS 2809-34401 (3 TURNS THROUGH BEAD)

Fig. 71-3

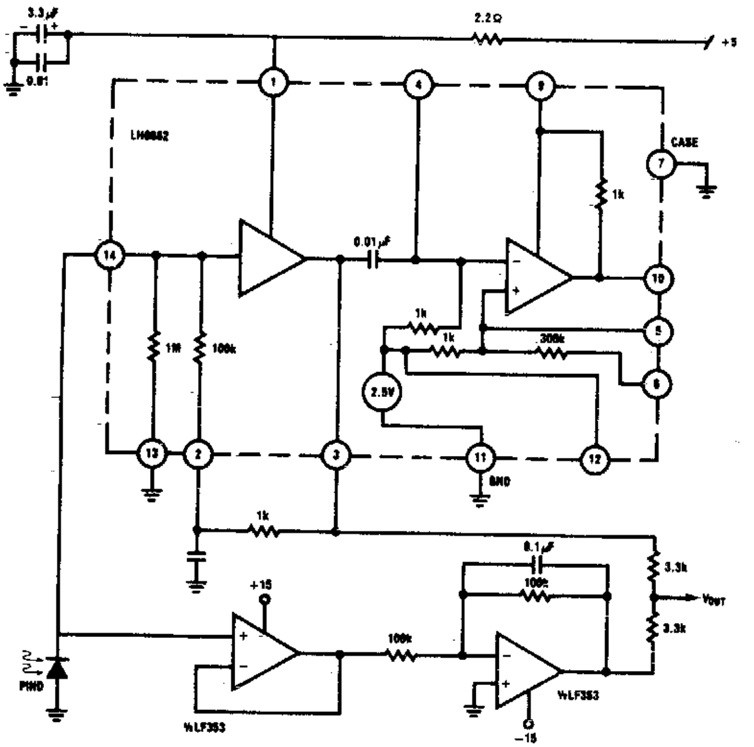
FM STEREO DEMODULATION SYSTEM



V⁺ = 12V
C10
0.1μF
C9
0.0025
COMPOSITE INPUT
1
16
C7
0.033
2
15
R4
21K
R1
3.9K
C6
2μF
C1
0.022
3
R3
3.3K
14
C8
0.22
C4
390pF
R5
5K
VCO ADJUST
100 mA STEREO LAMP
LEFT OUT
4
LM1800
13
C3
0.47μF
R2
3.9K
RIGHT OUT
5
12
C2
0.022
6
11
19 kHz MONITOR PIN
7
10
C5
0.33μF
8
9
TOP VIEW

Fig. 71-4

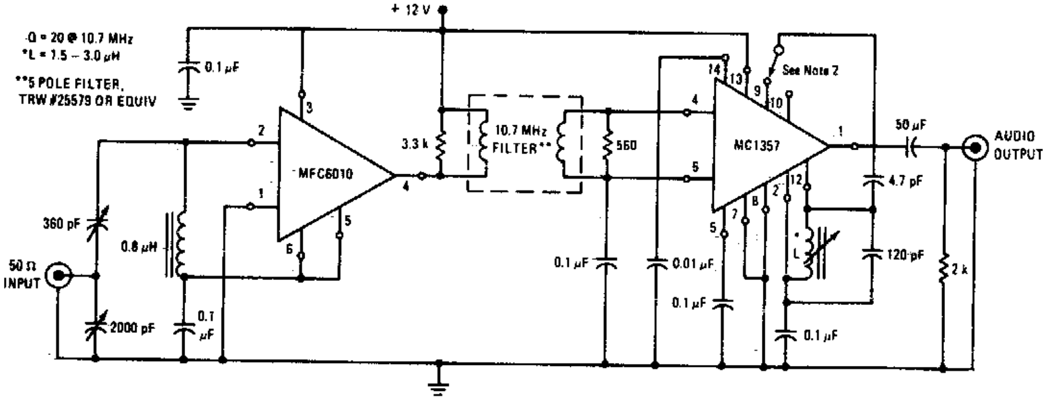
ANALOG RECEIVER (LOW TEMPERATURE DRIFT)



3.3 µF
0.01
2.2 Ω
+5
LH0082
CASE
1
4
9
7
1k
0.01 µF
10
5
6
1k
1k
300k
1M
100k
2.5V
14
13
2
3
11
12
GND
1k
0.1 µF
100k
3.3k
VOUT
3.3k
+15
100k
PIND
½LF353
½LF353
−15

Fig. 71-5

FM RADIO



+12 V
0.1 µF
3
2
1
MFC6010
4
5
6
360 pF
0.8 µH
50 Ω INPUT
2000 pF
0.1 µF
3.3 k
10.7 MHz FILTER**
560
0.1 µF
0.01 µF
0.1 µF
14
13
9
10
See Note 2
4
5
MC1357
1
7
8
2
12
50 µF
AUDIO OUTPUT
4.7 pF
L
120 pF
2 k
0.1 µF

·Q = 20 @ 10.7 MHz
*L = 1.5 – 3.0 µH

**5 POLE FILTER,
TRW #25579 OR EQUIV

Fig. 71-6

SIMPLE LF CONVERTER

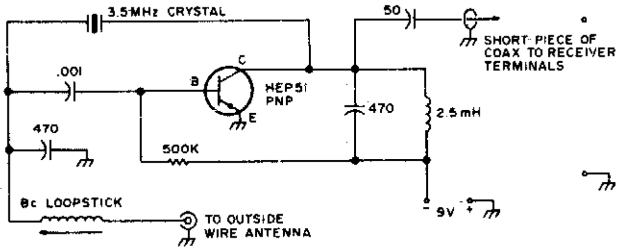
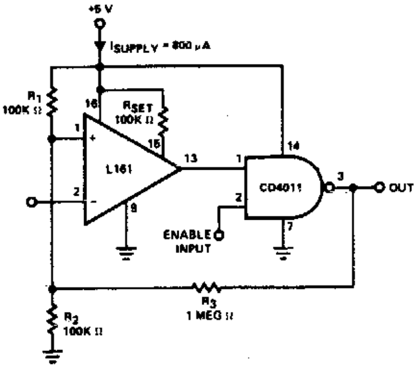

3.5MHz CRYSTAL
50
SHORT PIECE OF COAX TO RECEIVER TERMINALS
.001
B
C
E
HEP51 PNP
470
2.5mH
470
500K
8c LOOPSTICK
TO OUTSIDE WIRE ANTENNA
9V

Fig. 71-7

Circuit Notes

This converter allows coverage from 25 kHz up to 500 kHz. Use short coax from the converter to receiver antenna input. Tune the receiver to 3.5 MHz, peak for loudest crystal calibrator and tune your receiver higher in frequency to 3.6 MHz and you're tuning the 100 kHz range. 3.7 MHz puts you at 200 kHz, 3.8 MHz equals 300 kHz, 3.9 MHz yields 500 kHz, and 4.0 MHz gives you 500 kHz.

CMOS LINE RECEIVER

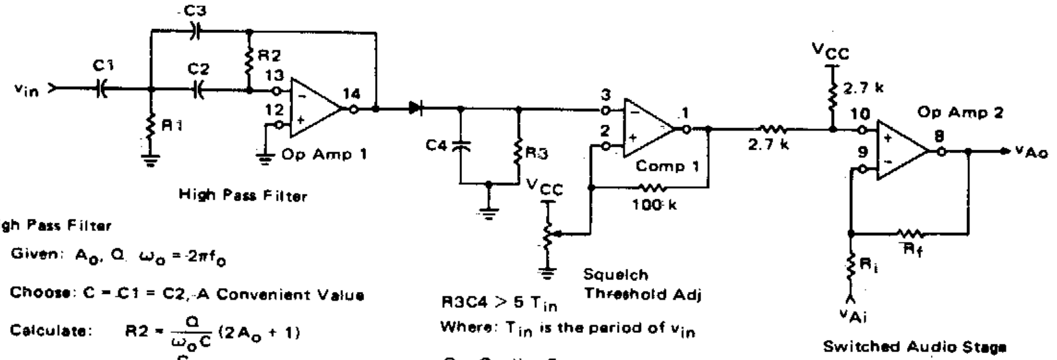

+5 V
ISUPPLY = 800 μA
R1 100K Ω
16
RSET 100K Ω
1
15
L161
13
2
9
1
14
CD4011
2
3
OUT
7
ENABLE INPUT
R3 1 MEG Ω
R2 100K Ω

Circuit Notes

The trip point is set half way between the supplies by R1 and R2; R3 provides over 200 mV of hysteresis to increase noise immunity. Maximum frequency of operation is about 300 kHz. If response to TTL levels is desired, change R2 to 39 K. The trip point is now centered at 1.4 V.

Fig. 71-8

SQUELCH CIRCUIT FOR AM OR FM


C3
C1
C2
R2
13
12
14
Vin
R1
Op Amp 1
High Pass Filter
C4
R3
Vcc
Squelch
Threshold Adj
3
2
1
Comp 1
100 k
2.7 k
Vcc
2.7 k
10
9
8
Op Amp 2
VAo
Rf
Ri
VAi
Switched Audio Stage

High Pass Filter

Given: A_o, Q, $\omega_o = 2\pi f_o$

Choose: C = C1 = C2, A Convenient Value

Calculate: $R2 = \frac{Q}{\omega_o C}(2A_o + 1)$

$C3 = \frac{C}{A_o}$

$R1 = \frac{A_o}{Q\omega_o C(2A_o + 1)}$

$R3C4 > 5\,T_{in}$

Where: T_{in} is the period of v_{in}

Q = Quality Factor

A_o = High Frequency Gain

ω_o = Break Frequency

Gain of Audio Stage

$A_{cl} = -\frac{R_f}{R_i}$

Fig. 71-9

VLF CONVERTER

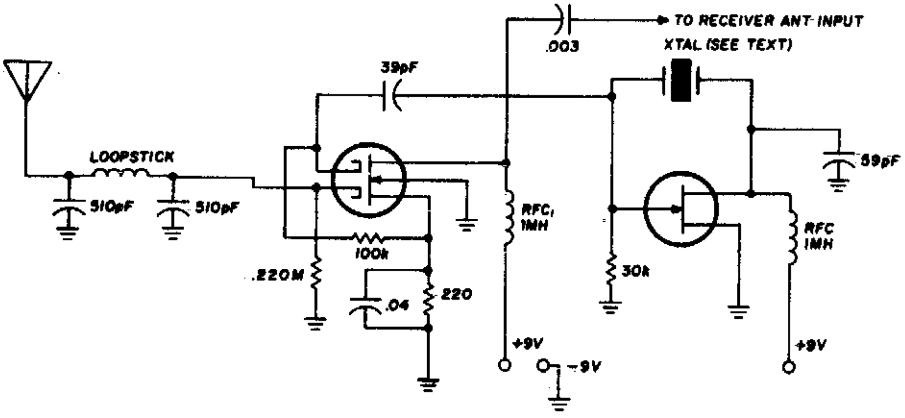

TO RECEIVER ANT INPUT
.003
XTAL (SEE TEXT)
39pF
59pF
LOOPSTICK
510pF
510pF
RFC1
1MH
100k
.220M
.04
220
30k
RFC
1MH
+9V
-9V
+9V

Fig. 71-10

Circuit Notes

This converter uses a low-pass filter instead of the usual tuned circuit so the only tuning required is with the receiver. The dual-gate MOSFET and FET used in the mixer and oscillator aren't critical. Any crystal having a frequency compatible with the receiver tuning range may be used. For example, with a 3500 kHz crystal, 3500 kHz on the receiver dial corresponds to zero kHz; 3600 to 100 kHz; 3700 to 200 kHz, etc. (At 3500 khz on the receiver all one can hear is the converter oscillator, and VLF signals start to come in about 20 kHz higher.)

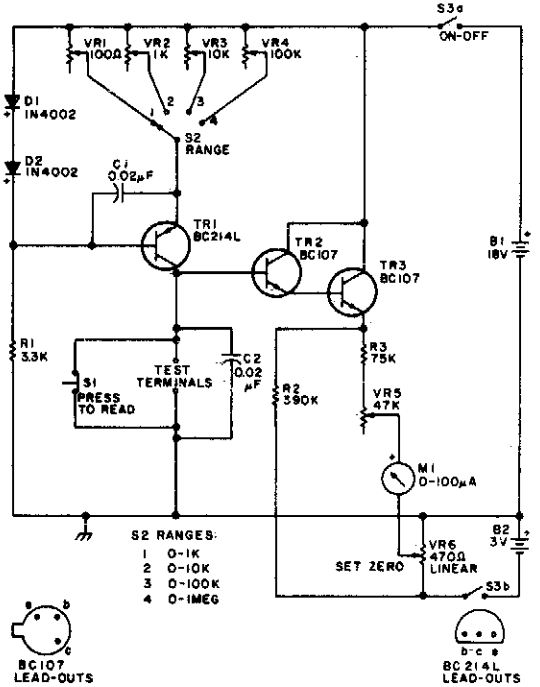
72

Resistance and Continuity Measuring Circuits

The sources of the following circuits are contained in the Sources section beginning on page 730. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

Linear Scale Ohmmeter
Ohmmeter
Low Parts Count Ratiometric Resistance
    Measurement
Audio Continuity Tester
Low Resistance Continuity Tester
"Buzz Box" Continuity and Coil Checker
Linear Scale Ohmmeter
Bridge Circuit

LINEAR SCALE OHMMETER


S3a
ON-OFF
VR1
100Ω
VR2
1K
VR3
10K
VR4
100K
1
2
3
4
S2
RANGE
D1
1N4002
D2
1N4002
C1
0.02μF
TR1
BC214L
TR2
BC107
TR3
BC107
B1
18V
R1
3.3K
S1
PRESS
TO READ
TEST
TERMINALS
C2
0.02
μF
R2
390K
R3
75K
VR5
47K
M1
0-100μA
B2
3V
VR6
470Ω
LINEAR
SET ZERO
S3b
S2 RANGES:
1 0-1K
2 0-10K
3 0-100K
4 0-1MEG
e
b
c
BC107
LEAD-OUTS
b c e
BC214L
LEAD-OUTS

Circuit Notes

This circuit is designed to provide accurate measurement and a linear resistance scale at the high end. The circuit has four ranges. Another meter with a current range of 10 μA to 10 mA and sensitivity of 10,000 ohms per volt is needed for setting up.

Fig. 72-1

OHMMETER

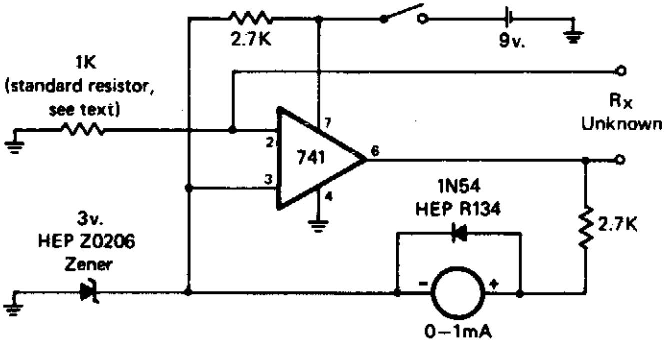

2.7K
9v.
1K
(standard resistor,
see text)
Rx
Unknown
2
3
7
6
4
741
1N54
HEP R134
2.7K
3v.
HEP Z0206
Zener
-
+
0–1mA

Fig. 72-2

Circuit Notes

This circuit has a linear reading scale, requires no calibration, and requires no zero adjustment. It may be made multirange by switching in different standard resistors.

LOW PARTS COUNT RATIOMETRIC RESISTANCE MEASUREMENT

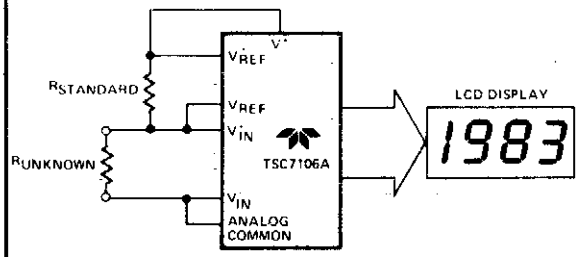

V+
VREF
RSTANDARD
VREF
VIN
RUNKNOWN
TSC7106A
VIN
ANALOG COMMON
LCD DISPLAY
1983

Fig. 72-3

Circuit Notes

The unknown resistance is put in series with a known standard and a current passed through the pair. The voltage developed across the unknown is applied to the input and the voltage across the known resistor applied to the reference input. If the unknown equals the standard, the display will read 1000. The displayed reading can be determined from the following expression:

$$\text{Displayed Reading} = \frac{R_{\text{unknown}}}{R_{\text{standard}}} \times 1000$$

The display will overrange for R_{unknown}, $\geqslant 2 \times R_{\text{standard}}$.

AUDIO CONTINUITY TESTER

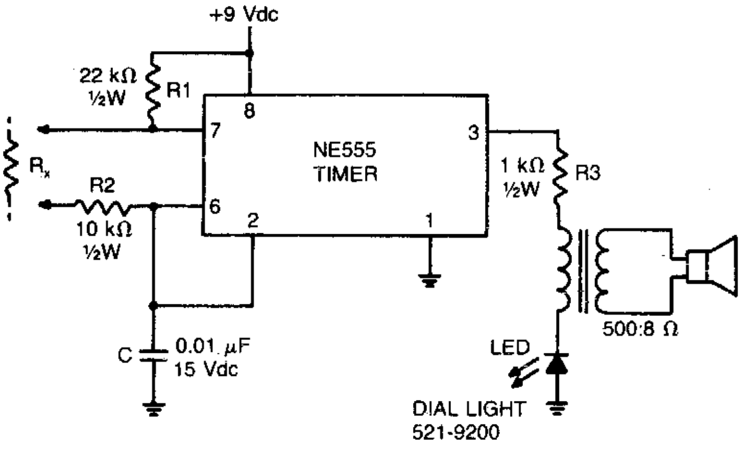

+9 Vdc
22 kΩ
½W
R1
8
7
3
NE555
TIMER
Rx
R2
6
2
1
10 kΩ
½W
1 kΩ
½W
R3
500:8 Ω
C
0.01 μF
15 Vdc
LED
DIAL LIGHT
521-9200

Fig. 72-4

Circuit Notes

This low-current audio continuity tester indicates the unknown resistance value by the frequency of audio tone. A high tone indicates a low resistance, and a tone of a few pulses per second indicates a resistance as high as 30 megohms.

LOW RESISTANCE CONTINUITY TESTER

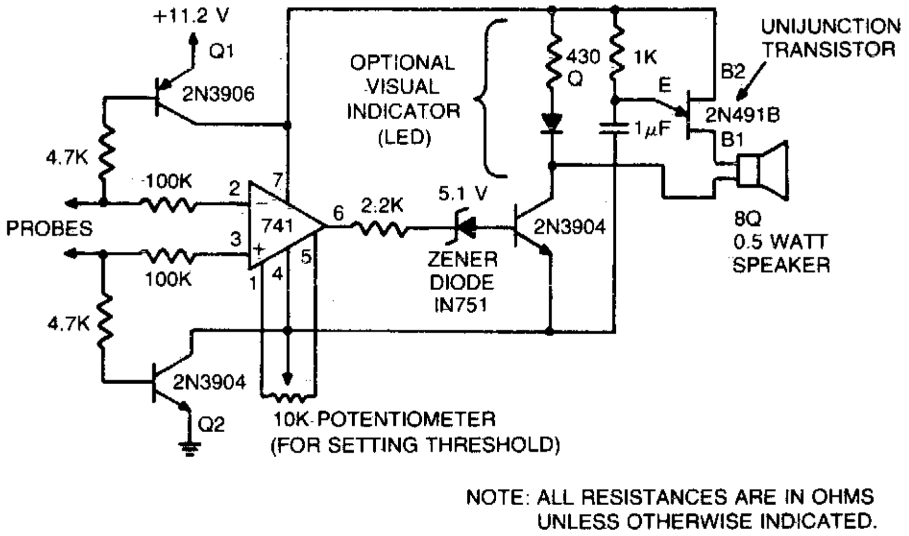

+11.2 V
Q1
2N3906
4.7K
100K
PROBES
100K
4.7K
2N3904
Q2
2
7
741
3
6
5
1
4
10K POTENTIOMETER
(FOR SETTING THRESHOLD)
OPTIONAL
VISUAL
INDICATOR
(LED)
2.2K
5.1 V
ZENER
DIODE
IN751
2N3904
430
Q
1K
1μF
E
B2
B1
UNIJUNCTION
TRANSISTOR
2N491B
8Q
0.5 WATT
SPEAKER
NOTE: ALL RESISTANCES ARE IN OHMS
UNLESS OTHERWISE INDICATED.

Fig. 72-5

Circuit Notes

This tester can be used to check IC printed circuit boards. Two 4.7 K resistors and the transistors connected to them prevent current flow through the operational amplifier until the probe circuit is completed. The zener diode in series with the operational amplifier output prevents audio oscillator operation until the positive output of the operational amplifier has sufficient amplitude.

"BUZZ BOX" CONTINUITY AND COIL CHECKER

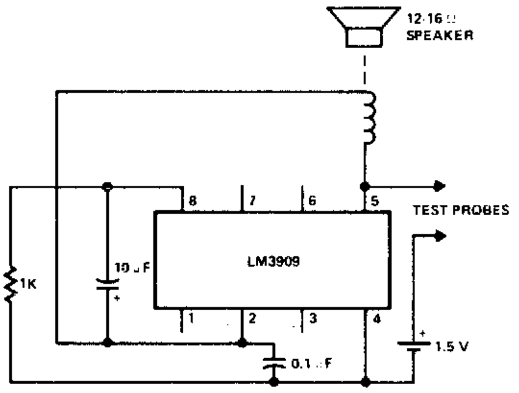
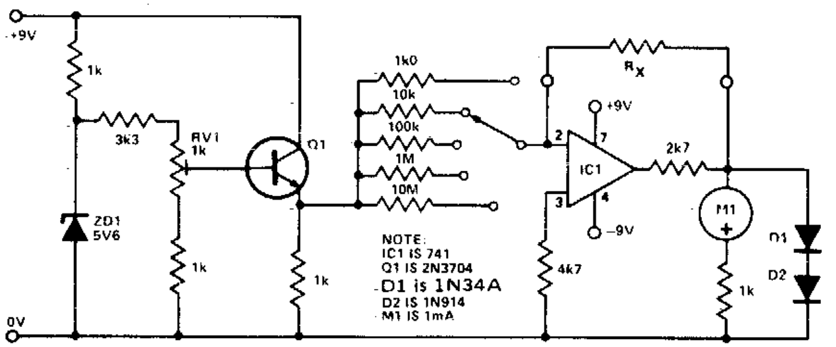

12-16 Ω
SPEAKER
8
7
6
5
TEST PROBES
LM3909
1
2
3
4
1K
10 μF
+
0.1 μF
+
1.5 V

Fig. 72-6

Circuit Notes

Differences between shorts, coils, and a few ohms of resitance can be heard.

LINEAR SCALE OHMMETER



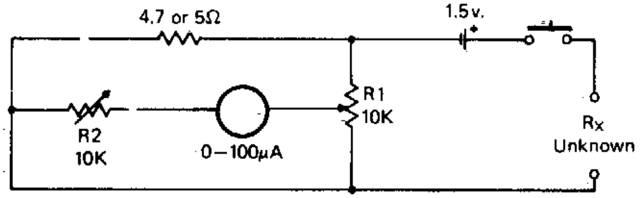
+9V
1k
3k3
RV1
1k
Q1
ZD1
5V6
1k
1k
0V
1k0
10k
100k
1M
10M
NOTE:
IC1 IS 741
Q1 IS 2N3704
D1 is 1N34A
D2 IS 1N914
M1 IS 1mA
R_X
+9V
2
7
IC1
3
4
−9V
2k7
4k7
M1
+
D1
D2
1k

Fig. 72-7

Circuit Notes

One preset resistor is used for all the ranges, simplifying the setting up. Diode clamping is included to prevent damage to the meter if the unknown resistor is higher than the range selected. When the meter has been assembled, a 10 K precision resistor is placed in the test position, R_x; the meter is set to the 10 K range and RV1 is adjusted for full scale deflection.

BRIDGE CIRCUIT



4.7 or 5Ω
1.5v.
+
R2
10K
0–100μA
R1
10K
R_X
Unknown

Fig. 72-8

Circuit Notes

For measurement of resistances from about 5 ohms down to about 1/10 ohm.

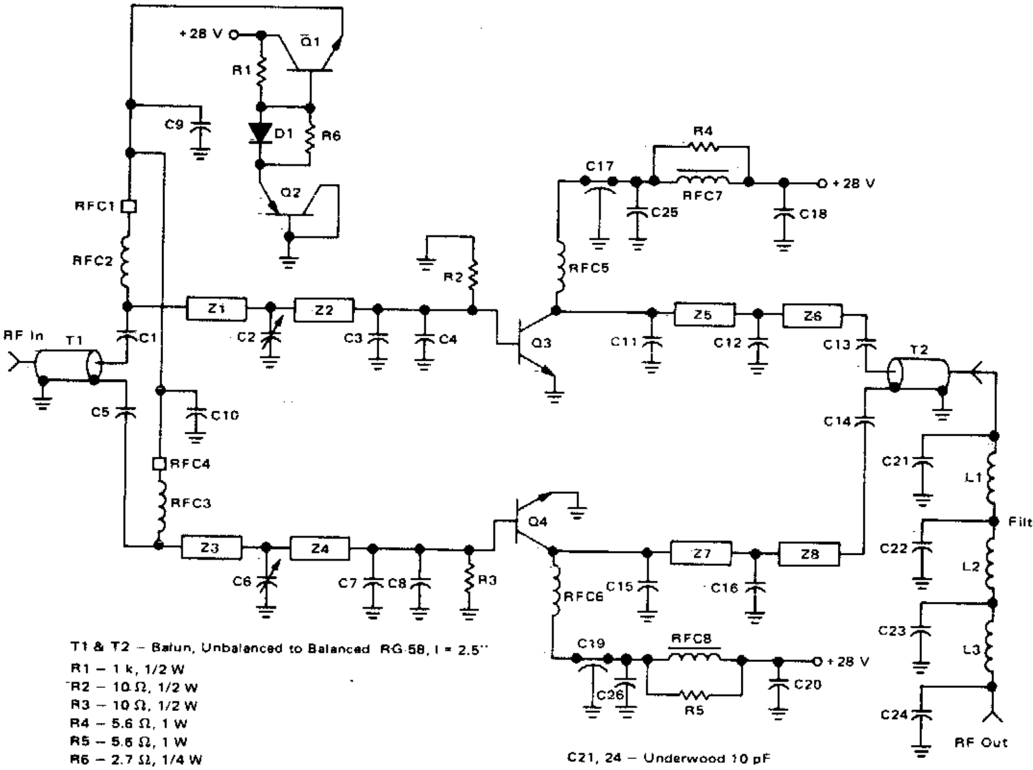
73

RF Amplifiers

The sources of the following circuits are contained in the Sources section beginning on page 730. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

100 W PEP 420-450 MHz Push-Pull Linear Amplifier
140 W (PEP) Amateur Radio Linear Amplifier (230 MHz)
160 W (PEP) Broadband Linear Amplifier
80 W (PEP) Broadband/Linear Amplifier
Single-Device, 80 W, 50 Ohm VHF Amplifier
600 W RF Power Amplifier
Wideband UHF Amplifier with High-Performance FETs
10 MHz Coaxial Line Driver
VHF Preamplifier
Shortwave FET Booster
Low-Noise 30 MHz Preamplifier
Low-Noise Broadband Amplifier
Two-Meter 10 Watt Power Amplifier
Two-Stage 60 MHz IF Amplifier
28 V Wideband Amplifier
200 MHz Cascode Amplifier
135-175 MHz Amplifier
200 MHz Cascode Amplifier
100 MHz and 400 MHz Neutralized Common Source Amplifier
Ultra High Frequency Amplifier
UHF Amplifier Inverting Gain of 2 with Lag-Lead Compensation
Transistorized Q-Multiplier for Use with IFs in the 1400 kHz Range
60 MHz Amplifier
30 MHz Amplfier
Two Meter Amplifier, 5 W Output
80 MHz Cascode Amplifier
200 MHz Neutralized Common Source Amplifier
450 MHz Common-Source Amplifier

100 W PEP 420-450 MHz PUSH-PULL LINEAR AMPLIFIER


+28 V
Q1
R1
C9
D1
R6
Q2
RFC1
RFC2
R4
C17
RFC7
+28 V
C25
C18
RFC5
R2
RF In
T1
C1
Z1
C2
Z2
C3
C4
Q3
C11
Z5
C12
Z6
C13
T2
C5
C10
C14
C21
L1
RFC4
RFC3
Q4
Filter
Z3
C6
Z4
C7
C8
R3
Z7
C15
C16
Z8
C22
L2
RFC6
C19
RFC8
+28 V
C23
L3
C26
R5
C20
C24
RF Out

T1 & T2 – Balun, Unbalanced to Balanced RG-58, l = 2.5″
R1 – 1 k, 1/2 W
R2 – 10 Ω, 1/2 W
R3 – 10 Ω, 1/2 W
R4 – 5.6 Ω, 1 W
R5 – 5.6 Ω, 1 W
R6 – 2.7 Ω, 1/4 W
Z1 & Z3 · Microstrip – W = 200 mils, l = 1.8″
Z2 & Z4 – Microstrip – W = 200 mils, l = 300 mils
Z5 & Z6 – Microstrip – W = 150 mils, l = 300 mils
Z6 & Z8 – Microstrip – W = 150 mils, l = 1.4″
RFC1, 4 – Ferroxcube Bead 56-590-65-3B
RFC2, 3 – 0.15 μH Cambion Molded Coil
RFC5, 6 – 1 Turn #20 Enameled Wire Wound on 5/16″ Bolt
RFC7, 8 – VK200 20/4B
C1, 3, 4, 5, 7, 8, 11, 15 – Underwood 40 pF
C12, 16 – Underwood 25 pF
C13, 14, 22, 23 – Underwood 15 pF
C9, 10, 18, 20 – 1 μF Tantalum

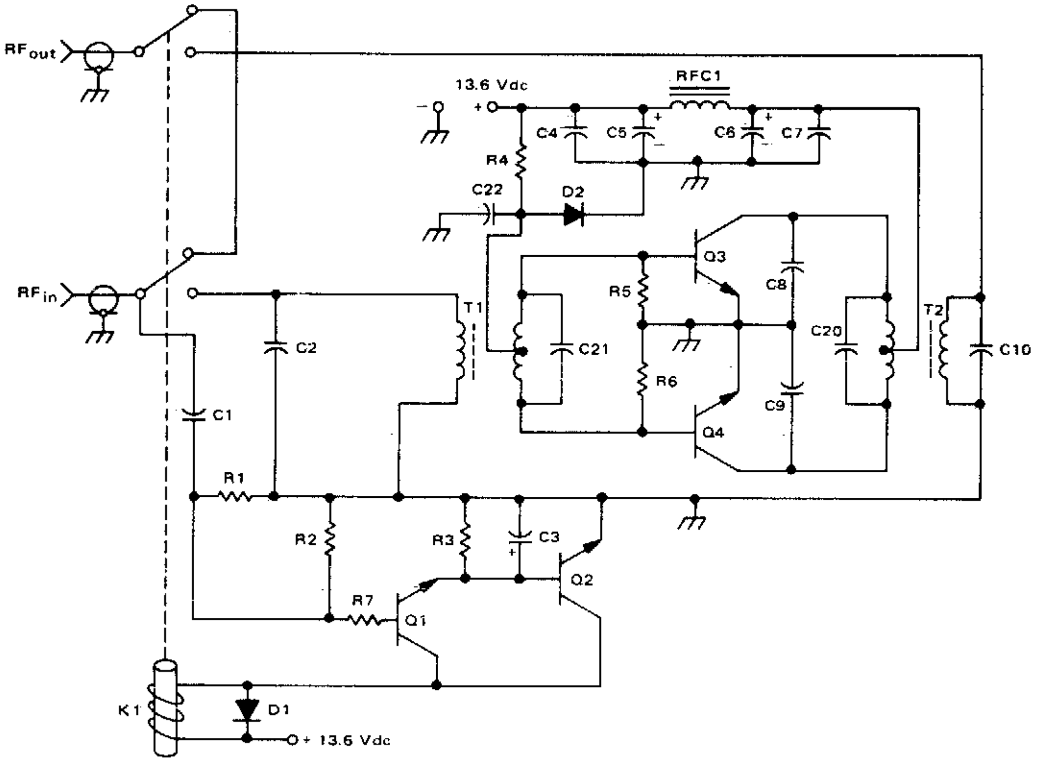
C21, 24 – Underwood 10 pF
C2, 6 – Arco 403
C17, 19 – Underwood J102, 1000 pF Feed Thru
C25, 26 – 0.1 μF, Erie Red Cap
L1 – 24 nH, #14 Wire, l = 1.2″
L2 – 12 nH, #14 Wire, l = 0.6″
L3 – 24 nH, #14 Wire, l = 1.2″
Board – G10, $\epsilon R \approx 5$, t = 0.062″, l = 8.0″, W = 4.0″
Q1 – 2N5192
Q2 – 2N5194
Q3 – MRF309
Q4 – MRF309
D1 – 1N4001

Fig. 73-1

Circuit Notes

This 100 watt linear amplifier may be constructed using two MRF309 transistors in push-pull, requiring only 16 watts drive from 420 to 450 MHz. Operating from a 28 volt supply, eight dB of power gain is achieved along with excellent practical performance featuring: maximum input SWR of 2:1, harmonic suppression more than—63 dB below 100 watts output, efficiency greater than 40%, circuit stability with a 3:1 collector mismatch at all phase angles.

140 W (PEP) AMATEUR RADIO LINEAR AMPLIFIER (2-30 MHz)



RF_{out}
RF_{in}
13.6 Vdc
RFC1
C4
C5
C6
C7
R4
C22
D2
Q3
R5
C8
T1
C2
C21
C20
T2
C10
R6
C9
C1
Q4
R1
R2
R3
C3
Q2
R7
Q1
K1
D1
+ 13.6 Vdc

C1 = 33 pF Dipped Mica
C2 = 18 pF Dipped Mica
C3 = 10 μF 35 Vdc for AM operation,
100 μF 35 Vdc for SSB operation.
C4 = .1 μF Erie
C5 = 10 μF 35 Vdc Electrolytic
C6 = 1 μF Tantalum
C7 = .001 μF Erie Disc
C8, 9 = 330 pF Dipped Mica
R1 = 100 kΩ 1/4 W Resistor
R2, 3 = 10 kΩ 1/4 W Resistor
R4 = 33 Ω 5 W Wire Wound Resistor
R5, 6 = 10 Ω 1/2 W Resistor

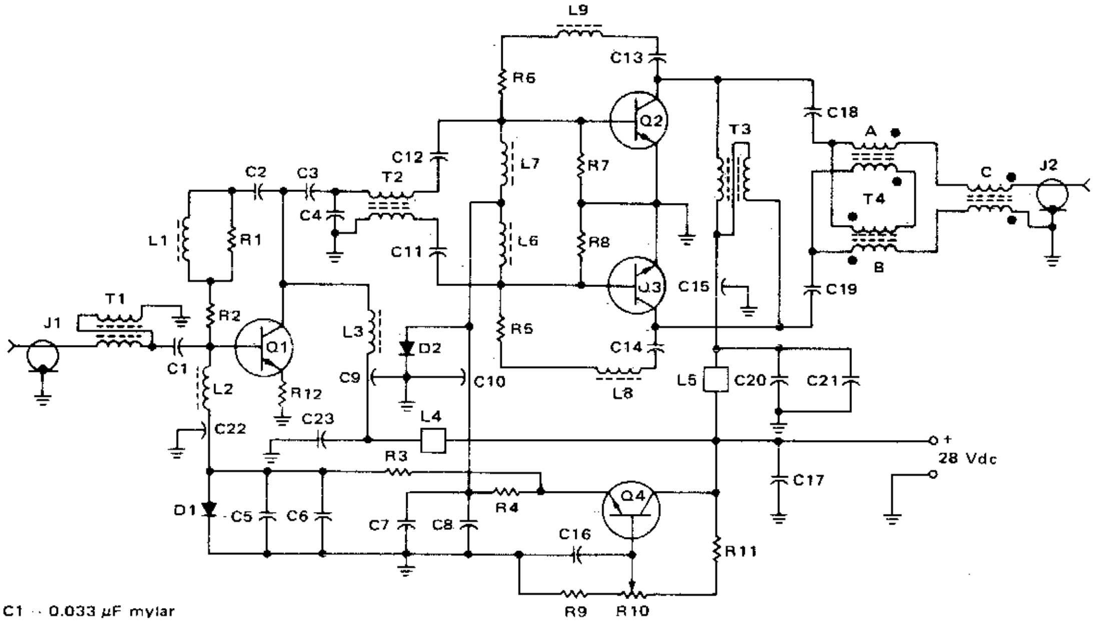
R7 = 100 Ω 1/4 W Resistor
RFC1 = 9 Ferroxcube Beads on #18 AWG Wire
D1 = 1N4001
D2 = 1N4997
Q1, Q2 = 2N4401
Q3, 4 = MRF454
T1, T2 = 16:1 Transformers
C20 = 910 pF Dipped Mica
C21 = 1100 pF Dipped Mica
C10 = 24 pF Dipped Mica
C22 = 500 μF 3 Vdc Electrolytic
K1 = Potter & Brumfield
KT11A 12 Vdc Relay or Equivalent

Fig. 73-2

Circuit Notes

This inexpensive, easy to construct amplifier uses two MRF454 devices. Specified at 80 W power output with 5 W of input drive, 30 MHz, and 12.5 Vdc.

160 W (PEP) BROADBAND LINEAR AMPLIFIER


L9
C13
R6
Q2
C18
T3
A
C12
L7
R7
J2
C2
C3
T2
C
T4
C4
L1
R1
L6
R8
C11
B
T1
R2
Q3
C15
C19
J1
L3
R5
Q1
D2
C14
C1
C9
C10
L5
C20
C21
L2
R12
L8
C23
L4
C22
+
28 Vdc
R3
C17
D1
C5
C6
R4
Q4
C7
C8
C16
R11
R9
R10

C1 -- 0.033 μF mylar
C2, C3 – 0.01 μF mylar
C4 – 620 pF dipped mica
C5, C7, C16 – 0.1 μF ceramic
C6 – 100 μF/15 V electrolytic
C8 -- 500 μF/6 V electrolytic
C9, C10, C15, C22 -- 1000 pF feed through
C11, C12 – 0.01 μF
C13, C14 – 0.015 μF mylar
C17 – 10 μF/35 V electrolytic
C18, C19, C21 -- Two 0.068 μF mylars in parallel
C20 – 0.1 μF disc ceramic
C23 - 0.1 μF disc ceramic
R1 – 220 Ω, 1/4 W carbon
R2 – 47 Ω, 1/2 W carbon
R3 – 820 Ω, 1 W wire W
R4 – 35 Ω, 5 W wire W
R5, R6 – Two 150 Ω, 1/2 W carbon in parallel
R7, R8 – 10 Ω, 1/2 W carbon
R9, R11 – 1 k, 1/2 W carbon
R10 – 1 k, 1/2 W potentiometer
R12 – 0.85 Ω (6 5.1 Ω or 4 3.3 Ω 1/4 W resistors in parallel, divided equally between both emitter leads)

T1 – 4:1 Transformer, 6 turns, 2 twisted pairs of #26 AWG enameled wire (8 twists per inch)

T2 -- 1:1 Balun, 6 turns, 2 twisted pairs of #24 AWG enameled wire (6 twists per inch)

T3 – Collector choke, 4 turns, 2 twisted pairs of #22 AWG enameled wire (6 twists per inch)

T4 – 1:4 Transformer Balun, A&B – 5 turns, 2 twisted pairs of #24, C – 8 turns, 1 twisted pair of #24 AWG enameled wire (All windings 6 twists per inch). (T4 – Indiana General F624-19Q1, – All others are Indiana General F627-8Q1 ferrite toroids or equivalent.)

PARTS LIST

L1 – .33 μH, molded choke
L2, L6, L7 – 10 μH, molded choke
L3 – 1.8 μH (Ohmite 2-144)
L4, L5 – 3 ferrite beads each
L8, L9 – .22 μH, molded choke

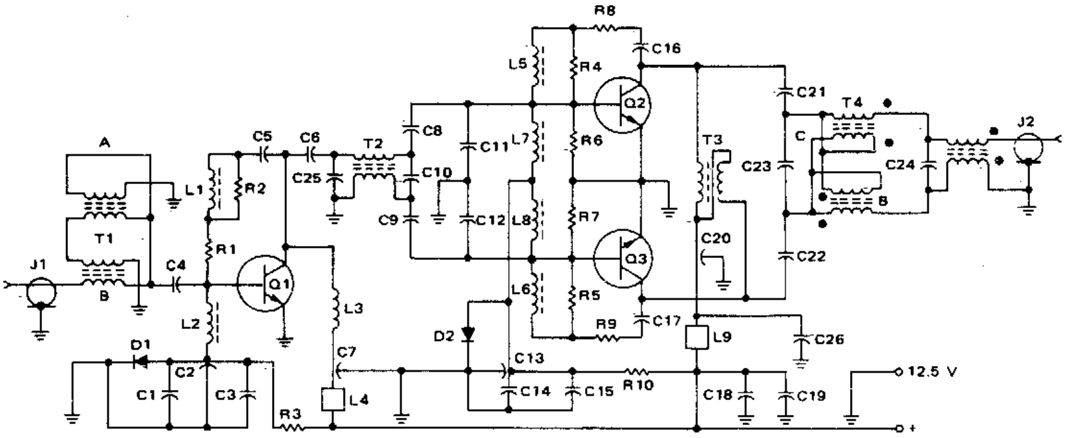
Q1 -- 2N6370
Q2, Q3 – 2N5942
Q4 – 2N5190

D1 – 1N4001
D2 – 1N4997

J1, J2 – BNC connectors

Fig. 73-3

80 W (PEP) BROADBAND/LINEAR AMPLIFIER


R8
C16
L5
R4
Q2
C21
T4
J2
A
C5
C6
T2
C8
C11
L7
R6
T3
C
C23
C24
L1
R2
C25
C10
C9
C12
L8
R7
B
T1
C20
J1
C4
R1
Q1
Q3
C22
B
L6
R5
L3
L2
D1
C7
D2
R9
C17
L9
C26
C13
C2
C1
C3
L4
C14
C15
R10
C18
C19
12.5 V
R3
+

C1, C14, C18 – 0.1 μF ceramic.
C2, C7, C13, C20 – 0.001 μF feed through.
C3 – 100 μF/3V.
C4, C6 – 0.033 μF mylar
C5 – 0.0047 μF mylar.
C8, C9 – 0.015 and 0.033 μF mylars in parallel.
C10 – 470 pF mica.
C11, C12 – 560 pF mica.
C15 – 1000 μF/3 V
C16, C17 – 0.015 μF mylar
C19 – 10 pF 15 V
C21, C22 – two 0.068 μF mylars in parallel.
C23 – 330 pF mica
C24 – 39 pF mica
C25 – 680 pF mica
C26 – .01 μF ceramic

R1, R6, R7 – 10 Ω, 1/2 W carbon.
R2 – 51 Ω, 1/2 W carbon
R3 – 240 Ω, 1 wire W
R4, R5 – 18 Ω, 1 W carbon
R8, R9 – 27 Ω, 2 W carbon
R10 – 33 Ω, 6 W wire W

L1 – 0.22 μh molded choke
L2, L7, L8 – 10 μh molded choke
L5, L6 – 0.15 μh
L3 – 25 t, #26 wire, wound on a 100 Ω, 2 W resistor. (1.0 μh)
L4, L9 – 3 ferrite beads each.

T1 – 2 twisted pairs of #26 wire, 8 twists per inch. A = 4 turns, B = 8 turns. Core- -Stackpole 57-9322-11, Indiana General F627-8Q1 or equivalent
T2 – 2 twisted pairs of #24 wire, 8 twists per inch, 6 turns. (Core as above.)
T3 – 2 twisted pairs of #20 wire, 6 twists per inch, 4 turns. (Core as above.)
T4 – A and B = 2 twisted pairs of #24 wire, 8 twists per inch. 5 turns each. C = 1 twisted pair of #24 wire, 8 turns. Core - -Stackpole 57-9074-11, Indiana General F624-19Q1 or equivalent.

Q1 – 2N6367

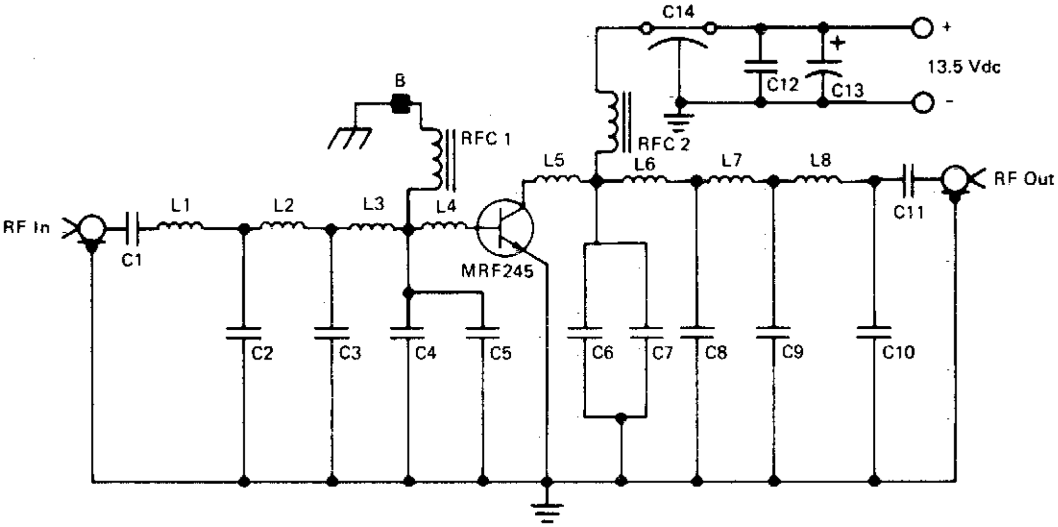
Q2, Q3 – 2N6368

D1 – 1N4001
D2 – 1N4997

J1, J2 – BNC connectors

Fig. 73-4

SINGLE-DEVICE, 80 W, 50 Ohm VHF AMPLIFIER


C14
+
13.5 Vdc
C12
C13
-
B
RFC 1
RFC 2
L5
L6
L7
L8
RF Out
RF In
L1
L2
L3
L4
C11
C1
MRF245
C2
C3
C4
C5
C6
C7
C8
C9
C10

C1, 11 – 500 pF Dipped mica
C2, 9 – 10 pF UNELCO
C3 – 60 pF UNELCO
C4, 5 – 250 pF UNELCO
C6, 7 – 250 pF UNELCO
C8 – 80 pF UNELCO
C10 – 40 pF UNELCO
C12 – 0.1 µF Erie Redcap
C13 – 1 µF Tantalum
C14 – 680 pF Allen Bradley Feed-Thru

RFC 1 – 0.15 µH Molded choke
RFC 2 – 10 T NO. 18 AWG Enameled Wire, 1/4" I.D.
B – Ferroxcube Bead 56-590-65, 3 Beads

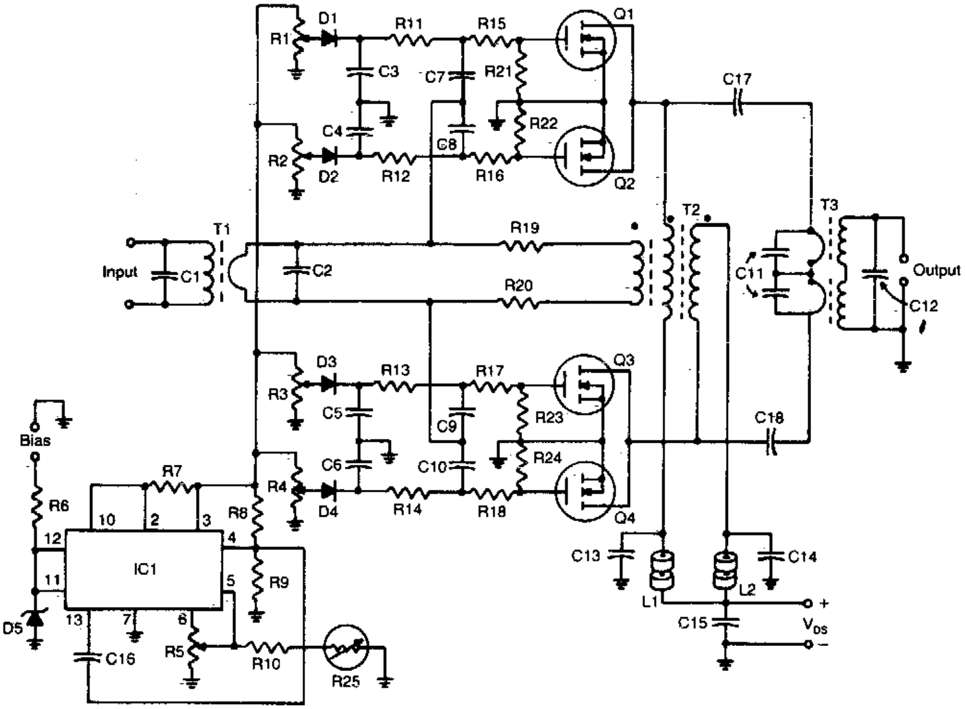
L1 – 1.2 X 0.3 cm Airline Inductor
L2 – 3.5 X 0.3 cm Airline Inductor
L3 – 4.0 X 0.3 cm Airline Inductor
L4, L5 – 0.3 X 0.3 cm Airline Inductor
L6 – 2.7 X 0.3 cm Airline Inductor
L7 – 0.8 X 0.3 cm Airline Inductor
L8 – 3.0 X 0.3 cm Airline Inductor
Board: G10, $\epsilon_r \approx 5$, t = 0.16 cm, 57 gm, Copper-Clad
connectors = BNC

Fig. 73-5

Circuit Notes

The amplifier uses a single MRF245 and provides 80 W with 9.4 dB gain across the 143 to 156 MHz band.

600 W RF POWER AMPLIFIER


D1
R11
R15
Q1
R1
C3
C7
R21
C17
C4
R22
C8
R2
D2
R12
R16
Q2
T1
R19
T2
T3
Input
C1
C2
C11
Output
R20
C12
D3
R13
R17
Q3
R3
C5
R23
C18
C9
Bias
C6
C10
R24
R7
R4
R6
10
2
3
R8
D4
R14
R18
Q4
12
4
IC1
C13
C14
11
5
R9
L1
L2
+
V_{DS}
D5
13
7
6
C15
−
C16
R5
R10
R25

R1-R5—10 kΩ trimpot
R6—1.0 kΩ/1.0W
R7—10 Ω
R8—2.0 kΩ
R9,R21-R24—10 kΩ
R10—8.2 kΩ
R11-R14—100 Ω
R15-R18—1.0 Ω
R19-R20—10 Ω/2.0 W Carbon
R25—thermistor, 10 kΩ (25°C), 2.5 kΩ (75°C)
C1—not used
C2—820 pF ceramic chip
C3-C6, C13,C14—0.1 μF ceramic
C7-C10—0.1 μF ceramic chip
C11—1200 pF each, 680 pF mica in parallel with an Arco 469 variable or three or more smaller value mica capacitors in parallel
C12—not used
C15—10 μF, 100 V electrolytic
C16—1000 pF ceramic
C17,C18—two 0.1 μF, 100 V ceramic each, (ATC 200/823 or equivalent)
D1-D4—IN4148
D5—28 V zener, IN5362 or equivalent
L1,L2—Two Fair-Rite 2673021801 ferrite beads each or equivalent 4.0 μH
T1-T3—see text
Q1-Q4—MRF150
IC1—MC1723CP

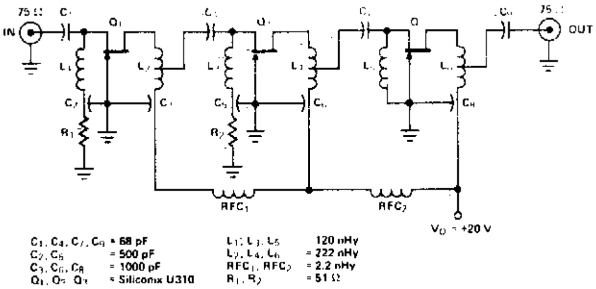
All resistors are 0.5W carbon or metal film unless otherwise designated.

Fig. 73-6

Circuit Notes

A unique push-pull parallel circuit. It uses four MRF150 RF power FETs paralleled at relatively high power levels. Supply voltages of 40 to 50 Vdc can be used, depending on linearity requirements. The bias for each device is independently adjustable; therefore, no matching is required for the gate threshold voltages.

WIDEBAND UHF AMPLIFIER WITH HIGH-PERFORMANCE FETs



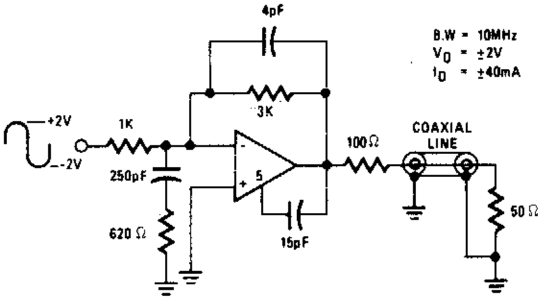
75 Ω
IN
OUT
RFC1
RFC2
V_D = +20 V
C_1, C_4, C_7, C_9 = 68 pF
C_2, C_5 = 500 pF
C_3, C_6, C_8 = 1000 pF
Q_1, Q_2, Q_3 = Siliconix U310
L_1, L_3, L_5 120 nHy
L_2, L_4, L_6 = 222 nHy
RFC_1, RFC_2 = 2.2 nHy
R_1, R_2 = 51 Ω

Circuit Notes

The amplifier circuit is designed for 225 MHz center frequency, 1 dB bandwidth of 50 MHz, low input VSWR in a 75-ohm system, and 24 dB gain. Three stages of U310 FETs are used in a straight forward design.

Fig. 73-7

10 MHz COAXIAL LINE DRIVER



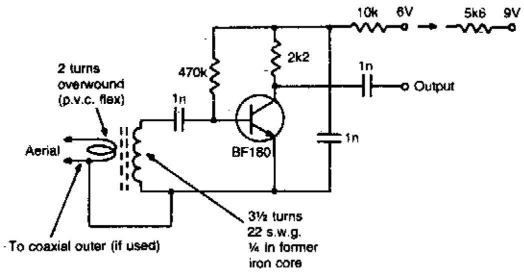
4pF
B.W = 10MHz
V_O = ±2V
I_O = ±40mA
3K
+2V
-2V
1K
250pF
620 Ω
5
15pF
100 Ω
COAXIAL LINE
50 Ω

Circuit Notes

The circuit will find excellent usage in high frequency line driving systems that require wide-power bandwidths at high output current levels. (IC=HA2530) The bandwidth of the circuit is limited only by the single pole response of the feedback components; namely $f(-3\text{ dB}) = \frac{1}{2}\pi R_f C_f$. As such, the response is flat with no peaking and yields minimum distortion.

Fig. 73-8

VHF PREAMPLIFIER



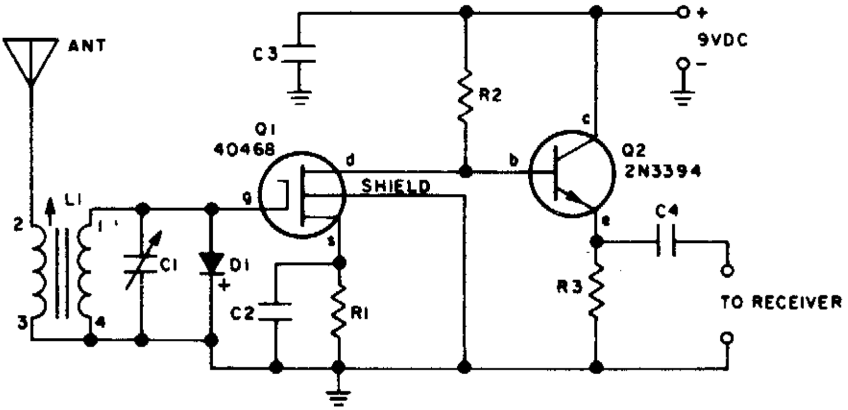
10k 6V
5k6 9V
2k2
470k
1n
Output
2 turns overwound (p.v.c. flex)
1n
Aerial
BF180
1n
3½ turns 22 s.w.g. ¼ in former iron core
To coaxial outer (if used)

Circuit Notes

This simple circuit gives 15 dB gain and can be mounted on 1 in²PCB. Coil data is given for 85 to 95 MHz. For other frequencies modify coil as required.

Fig. 73-9

SHORTWAVE FET BOOSTER


ANT
C3
+
9VDC
−
R2
Q1
40468
d
g
s
SHIELD
c
b
e
Q2
2N3394
C4
L1
2
3
1
4
C1
D1
+
C2
R1
R3
TO RECEIVER

PARTS LIST FOR SWL'S FET BOOSTER

C1—365-pF tuning capacitor
C2, C3—0.05-uF, 25-VDC capacitor
C4—470-pF, 25-VDC capacitor
D1—1N914 diode
L1—Antenna coil: 1.7-5.5 KHz use Miller B-5495A, 5.5-15 MHz use Miller C-5495A, 12-36 MHz use Miller D-5495-A
Q1—RCA 40468 FET transistor (Do not substitute)
Q2—2N3394 npn transistor
R1—470-ohm, ½-watt resistor
R2—2400-ohm, ½-watt resistor
R3—4700-ohm, ½-watt resistor

Fig. 73-10

Circuit Notes

This two transistor preselector provides up to 40 dB gain from 3.5 to 30 MHz. Q1 (MOSFET) is sensitive to static charges and must be handled with care.

LOW-NOISE 30 MHz PREAMPLIFIER

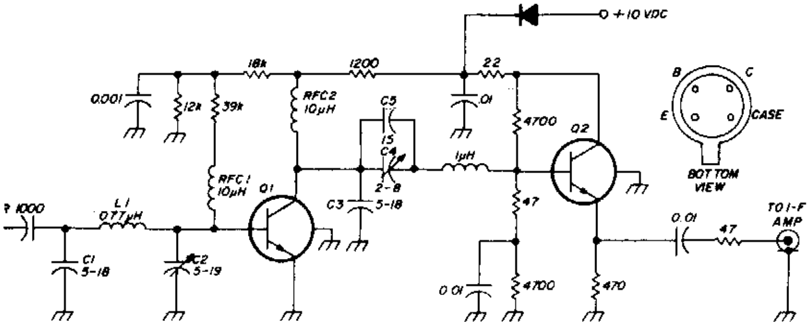

+10 VDC
18k
1200
22
0.001
12k
39k
RFC2
10μH
C5
15
C4
2-8
.01
4700
Q2
1μH
RFC1
10μH
Q1
C3
5-18
47
1000
L1
0.77μH
C1
5-18
C2
5-19
0.01
4700
470
0.01
47
TO I-F AMP
B
C
E
CASE
BOTTOM VIEW

Fig. 73-11

Circuit Notes

Low-noise preamplifier has a noise figure of 1.1 dB at 30 MHz and 3 dB bandwidth of 10 MHz. Gain is 19 dB. Total current drain with a +10 volt supply is 13 mA. All resistors are ¼ watt carbon; bypass capacitors are 50-volt ceramics.

LOW-NOISE BROADBAND AMPLIFIER

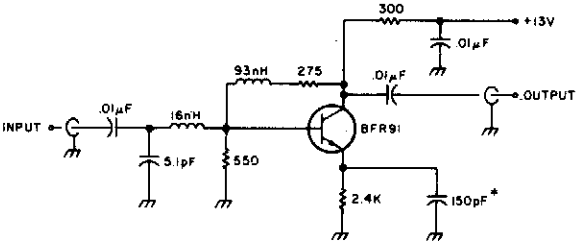

300
+13V
.01μF
93nH
275
.01μF
OUTPUT
.01μF
16nH
INPUT
BFR91
5.1pF
550
2.4K
150pF*

Fig. 73-12

Circuit Notes

The amplifier provides 10 dB of gain from 10-600 MHz and has a 1.5-to-1 match at 50 ohms. The BFR91 has a 1.5 dB noise figures at 500 MHz. The circuit requires 13.5 Vdc at about 13 mA. Keep the leads on the 150 pF emitter bypass capacitor as short as possible. The 16 nH coil is 2.5 turns of #26 enamel wire on the shank of a #40 drill. The 93 nH inductor is 10 turns of the same material.

TWO-METER 10 WATT POWER AMPLIFIER

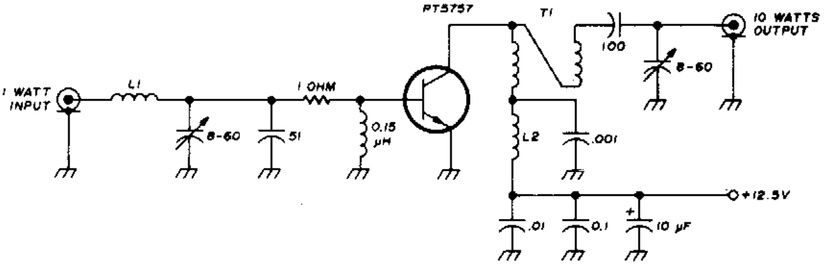
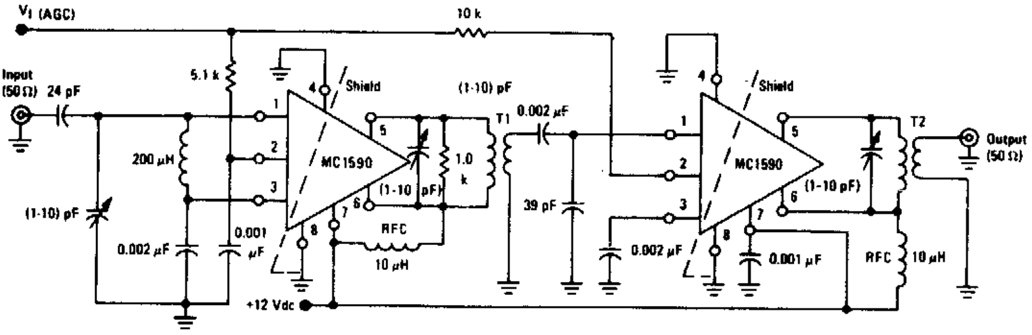

PT5757
T1
100
10 WATTS OUTPUT
8-60
1 WATT INPUT
L1
1 OHM
8-60
51
0.15 μH
L2
.001
+12.5V
.01
0.1
10 μF

Fig. 73-13

Circuit Notes

This 10-watt, 144-MHz power amplifier uses a TRW PT5757 transistor. L1 is 4 turns of no. 20 enameled, 3/32" ID; L2 is 10 turns of no. 20 enameled, 3/32" ID. Transformer T1 is a 4:1 transmission-line transformer made from a 3" length of twisted pair of no. 20 enameled wire.

TWO-STAGE 60 MHz IF AMPLIFIER
(POWER GAIN ≈ 80 dB, BW ≈ 1.5 MHz)



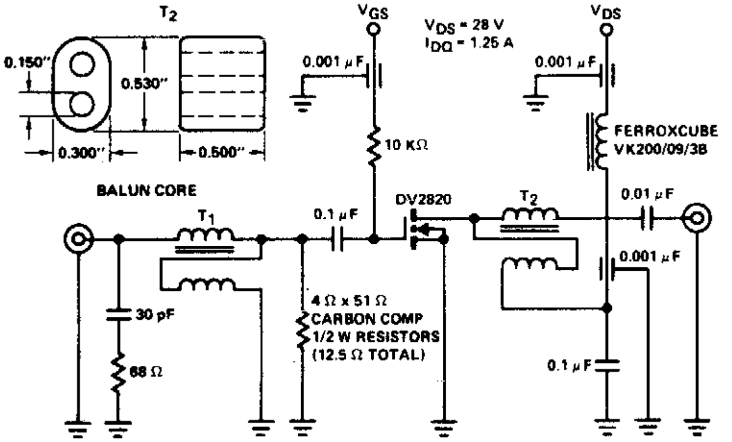
V_I (AGC)
10 k
Input
(50 Ω)
24 pF
5.1 k
Shield
MC1590
(1-10) pF
T1
0.002 μF
1.0 k
200 μH
(1-10) pF
(1-10 pF)
0.002 μF
0.001 μF
RFC
10 μH
39 pF
+12 Vdc
Shield
MC1590
T2
Output
(50 Ω)
(1-10 pF)
0.002 μF
0.001 μF
RFC
10 μH
1
2
3
4
5
6
7
8

T1: Primary Winding = 15 Turns, #22 AWG Wire, 1/4" ID Air Core
Secondary Winding = 4 Turns, #22 AWG Wire,
Coefficient of Coupling ≈ 1.0

T2: Primary Winding = 10 Turns, #22 AWG Wire, 1/4" ID Air Core
Secondary Winding = 2 Turns, #22 AWG Wire,
Coefficient of Coupling ≈ 1.0

Fig. 73-14

28 V WIDEBAND AMPLIFIER (3 to 100 MHz)



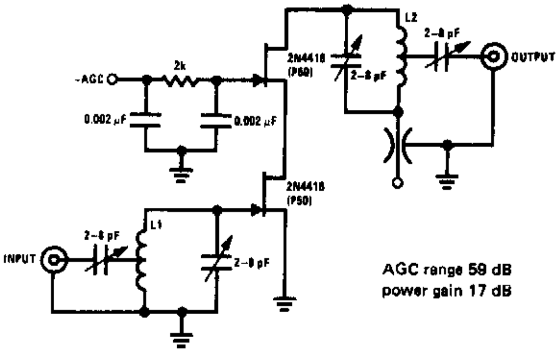
T_2
0.150"
0.530"
0.300"
0.500"
BALUN CORE
V_{GS}
V_{DS} = 28 V
I_{DQ} = 1.25 A
V_{DS}
0.001 μF
0.001 μF
10 KΩ
FERROXCUBE
VK200/09/3B
T_1
0.1 μF
DV2820
T_2
0.01 μF
0.001 μF
30 pF
68 Ω
4 Ω x 51 Ω
CARBON COMP
1/2 W RESISTORS
(12.5 Ω TOTAL)
0.1 μF

Fig. 73-15

Parts List

T_1, 20 turns 30 Ω, #30 bifilar on micrometals T-50-6 Toroid
T_2, 1 turn of 2-50 Ω coax cables in parallel through 2 balun cores stackpole #57-9130 μo = 125

200 MHz CASCODE AMPLIFIER


-AGC
2k
0.002 µF
0.002 µF
2N4416 (P50)
2N4416 (P50)
2–8 pF
2–8 pF
L2
2–8 pF
OUTPUT
INPUT
2–8 pF
L1
2–8 pF
AGC range 59 dB
power gain 17 dB

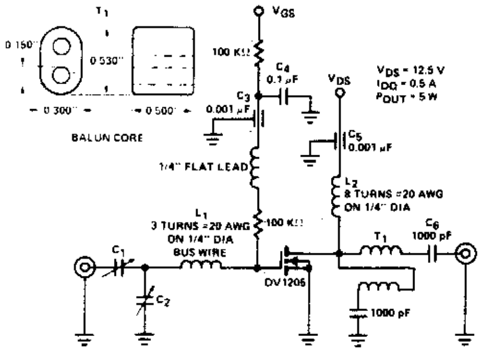
L1 = 0.07 µHy center tap
L2 = 0.07 µHy tap 1/4 up from ground

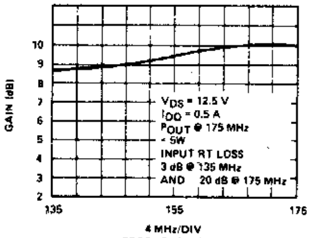
Circuit Notes

This 200 MHz JFET cascode circuit features low cross-modulation, large signal handling ability, no neutralization, and AGC controlled by biasing the upper cascode JFET. The only special requirement of this circuit is that Ioss of the upper unit must be greater than that of the lower unit.

Fig. 73-16

135-175 MHz AMPLIFIER


T1
0.150"
0.530"
0.300"
0.500"
BALUN CORE
VGS
100 KΩ
C4
0.1 µF
C3
0.001 µF
VDS
VDS = 12.5 V
IDQ = 0.5 A
POUT = 5 W
C5
0.001 µF
1/4" FLAT LEAD
L2
8 TURNS #20 AWG
ON 1/4" DIA
L1
3 TURNS #20 AWG
ON 1/4" DIA
BUS WIRE
100 KΩ
C1
C2
DV1205
T1
C6
1000 pF
1000 pF

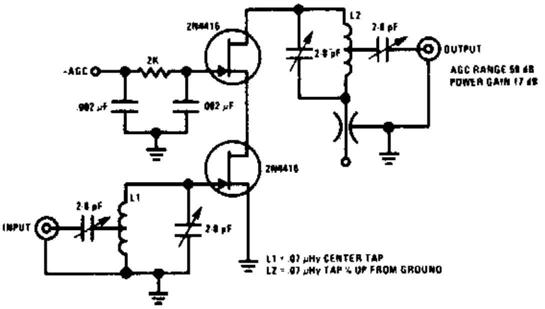


Parts List

C_1, C_2 ARCO #462, 2 to 80 pF, trimmer capacitors
L_1, 3 turns buss wire #20 AWG on 1/4" diameter
L_2, 8 turns #20 AWG on 1/4" diameter
T_1, 1 turn of 25 Ω coax on 2 balun cores.
Stackpole #57-0973 $\mu o = 35$.

Fig. 73-17

200 MHz CASCODE AMPLIFIER

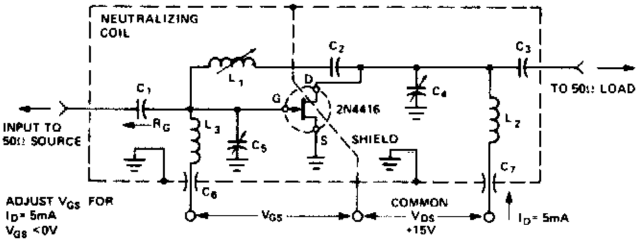

2N4416
-AGC
2K
.002 μF
.002 μF
2N4416
INPUT
2-8 pF
L1
2-8 pF
L2
2-8 pF
2-8 pF
OUTPUT
AGC RANGE 50 dB
POWER GAIN 17 dB
L1 = .07 μHy CENTER TAP
L2 = .07 μHy TAP ¼ UP FROM GROUND

Circuit Notes

This 200 MHz JFET cascode circuit features low cross-modulation, large signal handling ability, no neutralization, and AGC controlled by biasing the upper cascode JFET. The only special requirement of this circuit is that I_{DSS} of the upper unit must be greater than that of the lower unit.

Fig. 73-18

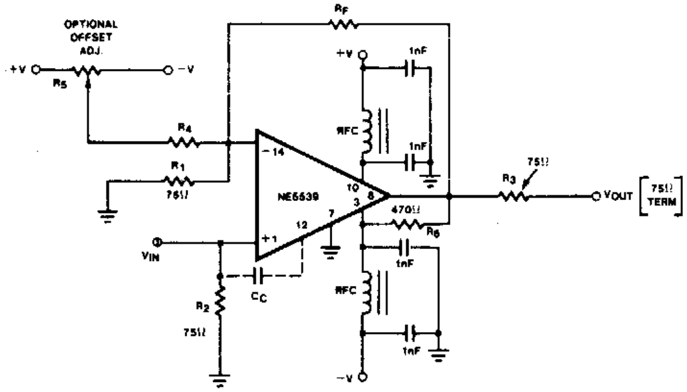
100 MHz AND 400 MHz NEUTRALIZED COMMON SOURCE AMPLIFIER


NEUTRALIZING COIL
C1
L1
C2
C3
D
G
2N4416
C4
TO 50Ω LOAD
INPUT TO 50Ω SOURCE
RG
L3
C5
S
SHIELD
L2
C6
C7
ADJUST VGS FOR
ID= 5mA
VGS <0V
VGS
COMMON
VDS
+15V
ID= 5mA

REFERENCE DESIGNATION	VALUE	
	100MHz	400MHz
C_1	7.0pF	1.8pF
C_2	1000pF	27pF
C_3	3.0pF	1.0pF
C_4	1.0–12pF	0.8–8pF
C_5	1.0–12pF	0.8–8pF
C_6	0.0015μF	0.001μF
C_7	0.0015μF	0.001μF
L_1	3.0μH	0.2μH
L_2	0.25μH	0.03μH
L_3	0.14μH	0.022μH
Typ NF	1.2dB	2.4dB
Typ G_{ps}	21dB	12dB

Fig. 73-19

ULTRA HIGH FREQUENCY AMPLIFIER


OPTIONAL OFFSET ADJ.
+V
R5
-V
RF
+V
1nF
RFC
1nF
R4
R1
75Ω
-14
NE5539
10
3
8
7
12
+1
VIN
CC
R2
75Ω
470Ω
R6
1nF
RFC
1nF
-V
75Ω
R3
VOUT
75Ω TERM

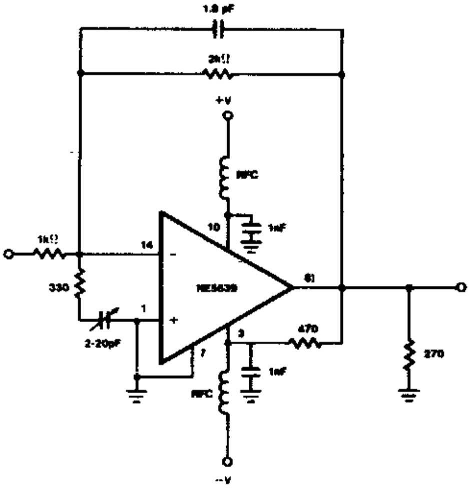
R_1 = 75Ω 5% CARBON
R_2 = 75Ω 5% CARBON
R_3 = 75Ω 5% CARBON
R_4 = 36K 5% CARBON

R_5 = 20K TRIMPOT (CERMET)
R_F = 1.5K (28dB GAIN)
R_6 = 470Ω 5% CARBON

RFC 3T # 28 BUSSWIRE ON
FERROXCUBE VK 200 09/3B CORE
BYPASS CAPACITORS
1nF CERAMIC
(MEPCO OR EQUIV.)

Fig. 73-20

UHF AMPLIFIER WITH INVERTING GAIN OF 2 AND LAG-LEAD COMPENSATION (GAIN BANDWIDTH PRODUCT 350 MHz)


1.8 pF
2kΩ
+V
RFC
10
1nF
1kΩ
14
−
NE5539
8
330
1
+
2-20pF
7
3
470
1nF
RFC
270
−V

NOTE
Resistors — 1/4 watt carbon.
RFC-3T #26 bus wire on Ferroxcube VK200 09/3B
wideband threaded core.

Fig. 73-21

TRANSISTORIZED Q-MULTIPLIER FOR USE WITH IFS IN THE 1400 kHz RANGE

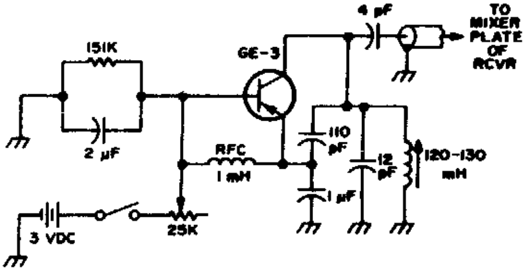
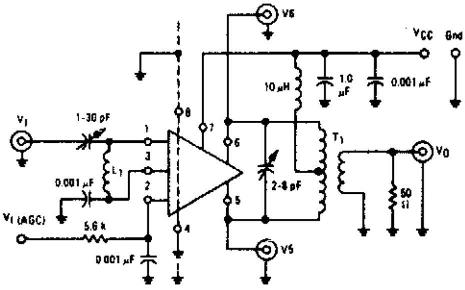

4 pF
TO
MIXER
PLATE
OF
RCVR
151K
GE-3
2 μF
RFC
1 mH
110
pF
12
pF
120-130
mH
1 μF
3 VDC
25K

Fig. 73-22

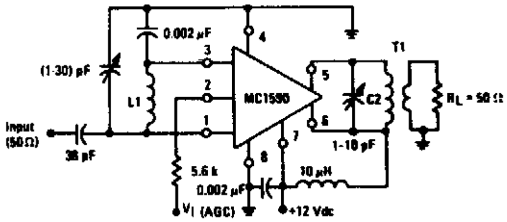
60 MHz AMPLIFIER


V6
VCC
Gnd
10 µH
1.0 µF
0.001 µF
VI
1-30 pF
8
7
1
3
2
6
5
4
T1
VO
L1
0.001 µF
2-8 pF
50 Ω
VI (AGC)
5.6 k
0.001 µF
V5

L1 7 Turns, # 22 AWG Wire
on 5/16" Dia. Form,
5/8" Long
T1: Close Wound Over 1/4" Form
Primary Winding = 16 Turns # 26 AWG, Center Tapped
Secondary Winding = 2 Turns # 26 AWG.

Fig. 73-23

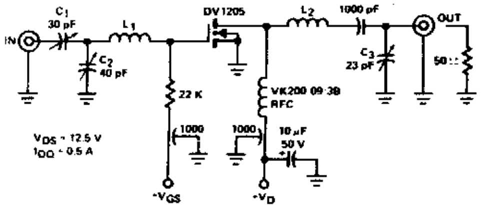
30 MHz AMPLIFIER (POWER GAIN = 50 dB, BW ≈ 1.0 MHz)


0.002 µF
4
3
T1
(1-30) pF
2
5
L1
MC1590
C2
RL = 50 Ω
1
6
Input
(50 Ω)
7
38 pF
1-10 pF
8
5.6 k
10 µH
0.002 µF
VI (AGC)
+12 Vdc

L1 = 12 Turns #22 AWG Wire on a Toroid Core,
(T37-6 Micro Metal or Equiv)
T1: Primary = 17 Turns #20 AWG Wire on a Toroid Core,
(T44-6 Micro Metal or Equiv)
Secondary = 2 Turns #20 AWG Wire

Fig. 73-24

TWO METER AMPLIFIER, 5 W OUTPUT


C1
30 pF
L1
DV1205
L2
1000 pF
OUT
IN
C2
40 pF
C3
23 pF
50 Ω
22 K
VK200 09-38
RFC
1000
1000
10 µF
50 V
V_{DS} = 12.5 V
I_{DQ} = 0.5 A
+V_{GS}
+V_D

Parts List

L_1, 60 nHy 4T #22 AWG close wound 0.125" I.D.
L_2, 54 nHy 3 1/2T #22 AWG close wound 0.125" I.D.
C_1, C_2, C_3, ARCO #462 5-80 pF

Fig. 73-25

80 MHz CASCODE AMPLIFIER

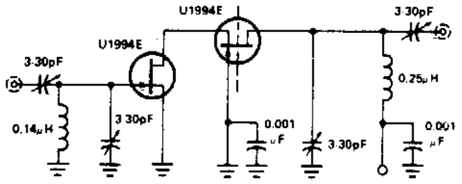
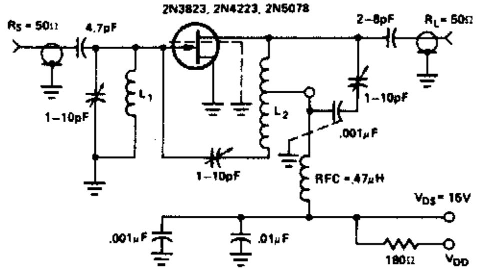

U1994E
3-30pF
U1994E
3-30pF
0.25µH
0.14µH
3-30pF
0.001 µF
3-30pF
0.001 µF

Fig. 73-26

200 MHz NEUTRALIZED COMMON SOURCE AMPLIFIER

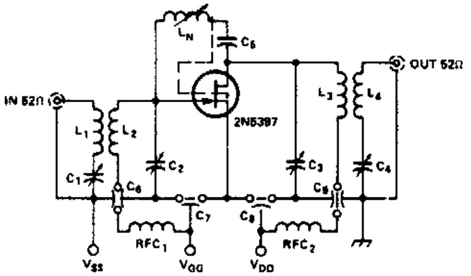

2N3823, 2N4223, 2N5078
R_S = 50Ω
4.7pF
1–10pF
L_1
L_2
1–10pF
2–8pF
R_L = 50Ω
1–10pF
.001μF
RFC = .47μH
V_{DS} = 15V
.001μF
.01μF
180Ω
V_{DD}

L_1 1-1/2 turns, #20 tinned wire, 1/4 ID, Length = 3/8"
L_2 3-1/2 turns, #18 tinned wire, 3/8" ID, Length = 1/2"
Tapped at 1-1/4 turns from drain

NF Typ 1.5dB
G_{ps} Typ 18dB
V_{DS} = +15V
V_{GS} = 0

Fig. 73-27

450 MHz COMMON-SOURCE AMPLIFIER


L_N
C_5
OUT 52Ω
IN 52Ω
L_1
L_2
2N5397
C_2
C_1
C_6
L_3
L_4
C_3
C_4
C_9
C_7
C_8
RFC_1
RFC_2
V_{SS}
V_{GG}
V_{DD}

C_{1-4} – 0.8 – 12pF Johanson type 2950
C_5 – 40pF DMS silver mica
C_{6-9} – 1000pF Allen-Bradley type FA5C
L_1 – 1.4" long; #22 enamel spaced 0.1" from L_2
L_2 – 1.1" long; #16 solid copper
L_3 – 1.3" long; #16 solid copper
L_4 – 1.4" long; #22 enamel spaced 0.3" from L_3
$RFC_{1,2}$ – 0.15μH Delevan type 1537-00
L_N – 3T, #22 enamel; 0.25" diam. ceramic form; aluminum slug, low loss

NF Typ 2.8dB
G_{ps} Typ 18dB
V_{DG} = +10V
I_D = 10mA

Fig. 73-28

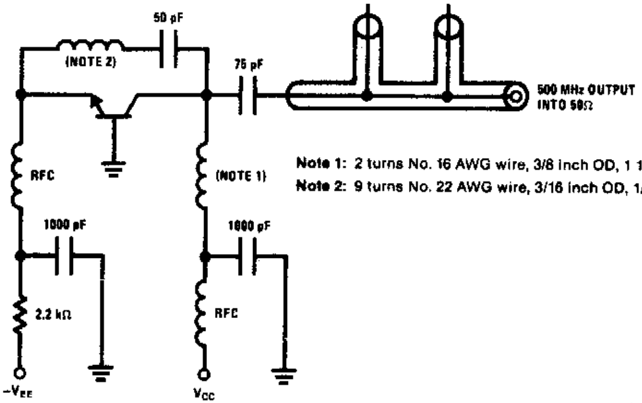
74

RF Oscillators

The sources of the following circuits are contained in the Sources section beginning on page 730. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

500 MHz Oscillator
Low Distortion Oscillator
400 MHz Oscillator
2 MHz Oscillator
1.0 MHz Oscillator
Hartley Oscillator
Colpitts Oscillator
RF Oscillator

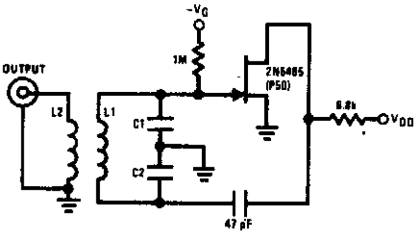
500 MHz OSCILLATOR


50 pF
(NOTE 2)
75 pF
500 MHz OUTPUT INTO 50Ω
RFC
(NOTE 1)
1000 pF
1000 pF
2.2 kΩ
RFC
$-V_{EE}$
V_{CC}

Note 1: 2 turns No. 16 AWG wire, 3/8 inch OD, 1 1/4 inch long.
Note 2: 9 turns No. 22 AWG wire, 3/16 inch OD, 1/2 inch long.

Fig. 74-1

LOW DISTORTION OSCILLATOR


$-V_G$
1M
OUTPUT
2N5485
(P50)
L2
L1
C1
C2
6.8k
V_{DD}
47 pF

20 MHz oscillator values

C1 ≅ 700 pF L1 = 1.3 μH
C2 = 75 pF L2 = 10T 3/8" dia 3/4" long
V_{DD} = 16V I_D = 1 mA

20 MHz oscillator performance

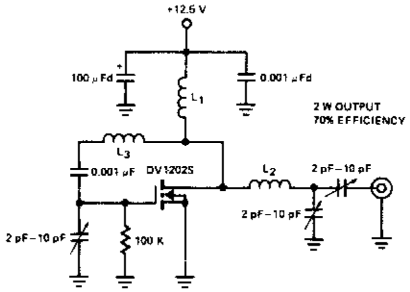
Low distortion 20 MHz osc
2nd harmonic – 60 dB
3rd harmonic > –70 dB

Fig. 74-2

Circuit Notes

The 2N5485 JFET is capable of oscillating in a circuit where harmonic distortion is very low. The JFET local oscillator is excellent when a low harmonic content is required for a good mixer circuit.

400 MHz OSCILLATOR


+12.5 V
100 μ Fd
L1
0.001 μ Fd
2 W OUTPUT
70% EFFICIENCY
L3
0.001 μF
DV1202S
L2
2 pF – 10 pF
2 pF – 10 pF
2 pF – 10 pF
100 K

Parts List

L_1 — 8 turns #22 closewound on 1/4" diameter
L_2 — 1/2 inch #16 wire
L_3 — 1 inch #16 wire

Fig. 74-3

1.0 MHz OSCILLATOR

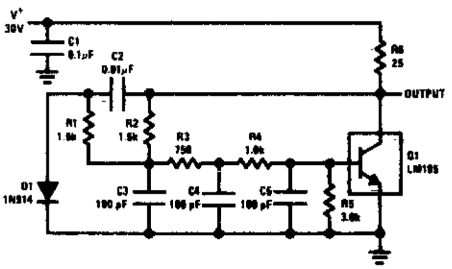

V+
30V
C1
0.1μF
C2
0.01μF
R6
25
OUTPUT
R1
1.5k
R2
1.5k
R3
750
R4
1.0k
Q1
LM195
D1
1N914
C3
100 pF
C4
100 pF
C5
100 pF
R5
3.0k

Fig. 74-5

2 MHz OSCILLATOR

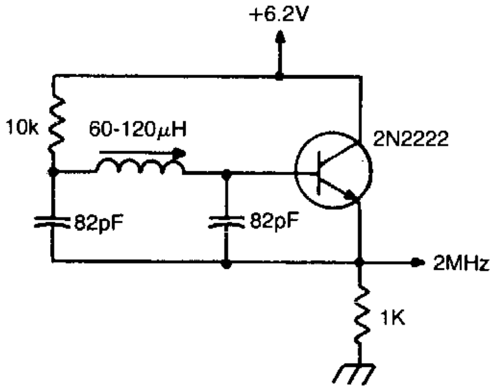

+6.2V
10k
60-120μH
2N2222
82pF
82pF
2MHz
1K

Fig. 74-4

Circuit Notes

Miller 9055 miniature slugtuned coil; all resistors 1/4W 5%; all caps min. 25 V ceramic.

HARTLEY OSCILLATOR

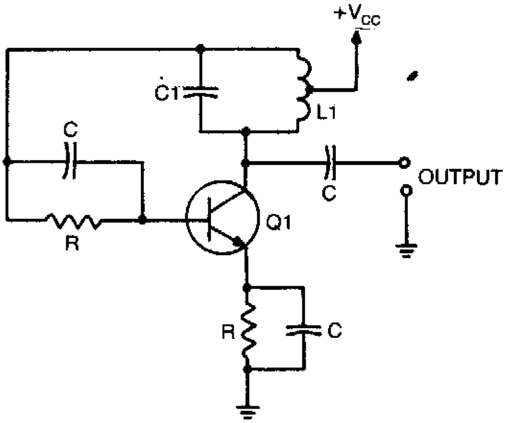
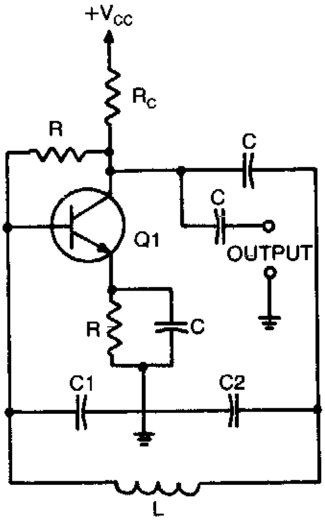

+V_CC
C1
L1
C
C
OUTPUT
R
Q1
R
C

Fig. 74-6

Circuit Notes

Resonant frequency is $\frac{1}{2} \pi \sqrt{L1C1}$.

COLPITTS OSCILLATOR

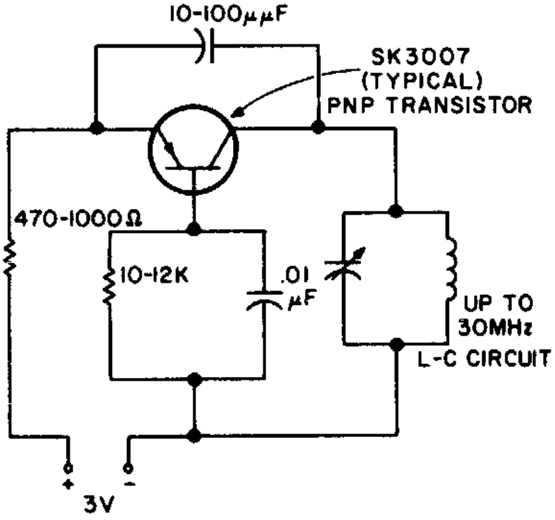

+V$_{CC}$
R$_C$
R
C
C
Q1
OUTPUT
R
C
C1
C2
L

Circuit Notes

When calculating its resonant frequency, use C1C2/C1+C2 for the total capacitance of the L-C circuit.

Fig. 74-7

RF OSCILLATOR


10-100μμF
SK3007
(TYPICAL)
PNP TRANSISTOR
470-1000Ω
10-12K
.01
μF
UP TO
30MHz
L-C CIRCUIT
+
-
3V

Circuit Notes

This rf oscillator is useful up to 30 MHz. An SK 3007 PNP transistor is recommended.

Fig. 74-8

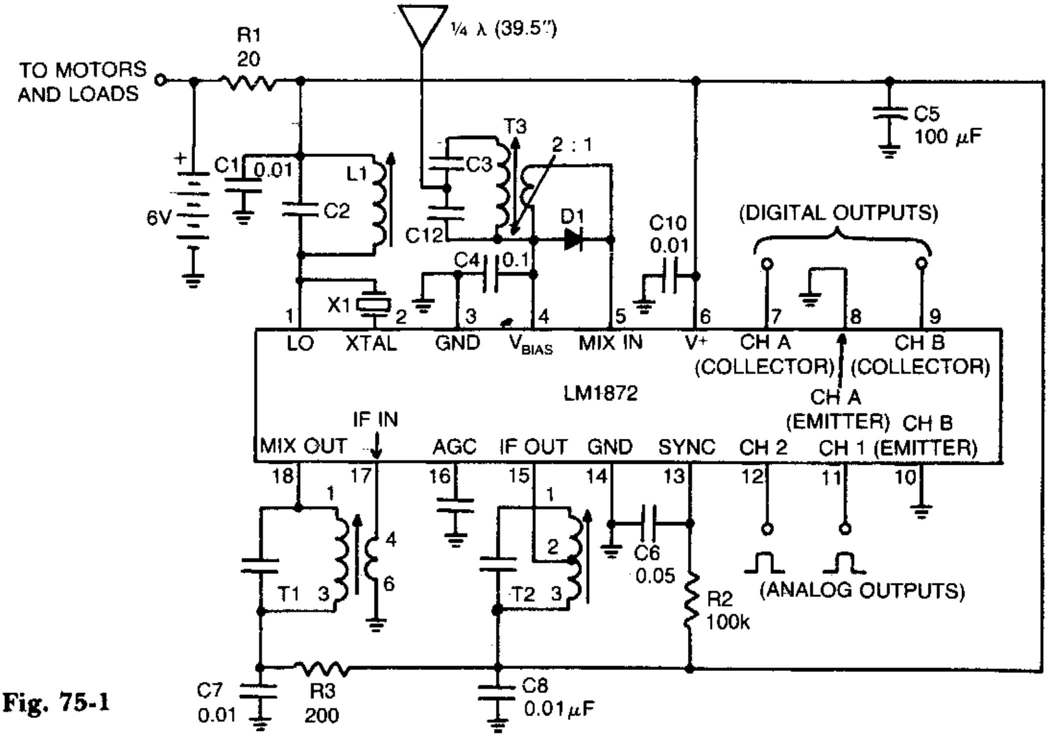
75

Remote Control Circuits

The sources of the following circuits are contained in the Sources section beginning on page 730. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

- Radio Control Receiver/Decoder
- Carrier Operated Relay
- Remote Control Servo System
- Tone-Actuated Relay
- Radio Control Motor Speed Controller
- Remote On-Off Switch
- Automatic Turn Off for TV Set

RADIO CONTROL RECEIVER/DECODER

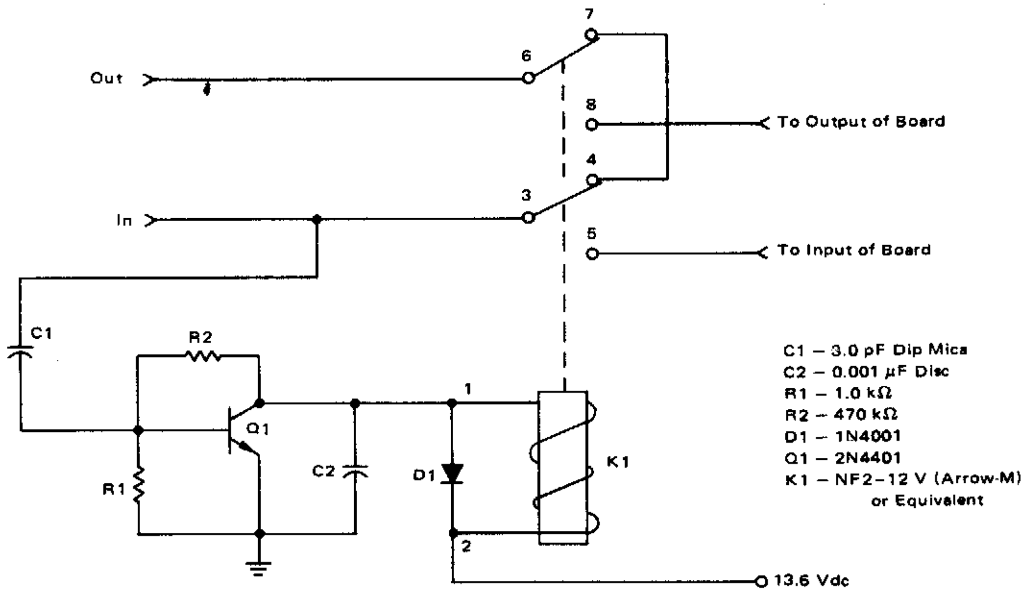


¼ λ (39.5")
R1
20
TO MOTORS AND LOADS
C5
100 μF
T3
2 : 1
C3
+
C1
0.01
L1
6V
C2
C12
D1
C10
0.01
(DIGITAL OUTPUTS)
C4
0.1
X1
1
2
3
4
5
6
7
8
9
LO
XTAL
GND
V_BIAS
MIX IN
V+
CH A (COLLECTOR)
CH B (COLLECTOR)
LM1872
CH A (EMITTER)
CH B
IF IN
MIX OUT
AGC
IF OUT
GND
SYNC
CH 2
CH 1 (EMITTER)
18
17
16
15
14
13
12
11
10
1
4
6
T1
3
1
2
T2
3
C6
0.05
R2
100k
(ANALOG OUTPUTS)
C7
0.01
R3
200
C8
0.01 μF

Fig. 75-1

R1 - Motor decoupling
R2 - Sync timer; $R2 = \frac{t_{SYNC}}{0.7\,C6}$, $R2 \leq 470k$
R3 - Mixer decoupling
C1 - LO bypass; optional
C2 - LO tank; C2 = 22 pF @ 72 MHz
C3 - Ant. input tank; C3 = 24 pF @ 72 MHz
C4 - V_{BIAS} bypass
C5 - Motor decoupling
C6 - Sync timer; $C6 = \frac{t_{SYNC}}{0.7\,R2}$, $C6 + 0.5\ \mu F$
C7 - Mixer decouple; $0.01\ \mu F \leq C7 \leq 1\ \mu F$
C8 - AGC
C9 - IF bypass; optional
C10 - V+ bypass; $0.01\ \mu F \leq C10 \leq 0.1\ \mu F$
C12 - Ant. input tank; C12 = 160 pF @ 72 MHz
L1 - LO coil
  Toko* 10k type (KENC) 4T; 0.2 μH @ 72 MHz
  L1 could be made a fixed coil, if desired.
T1 - 455 kHz mixer transformer
  Toko* 10 EZC type (RMC-502182), Qu = 110
  Pin 1-2, 82T; pin 2-3, 82T
  Pin 1-3, 164T; pin 4-6, 30T
T2 - 455 kHz IF transformer
  Toko* 10 EZC type (RMC-502503), Qu = 110
  Pin 1-2, 82T; pin 2-3, 8T
T3 - Ant. input transformer
  Toko 10k type (KENC), 4T sec. & 2T pri. of 0.2 μH@ 72 MHz
X1 - 5th overtone crystal, parallel-mode, 72 MHz
D1 - Electrostatic discharge (ESD) protection
* Toko America, Inc.
5520 West Touhy Ave.
Skokie, Ill. 60077
(312)677-3640 Tlx: 72-4372

CARRIER OPERATED RELAY

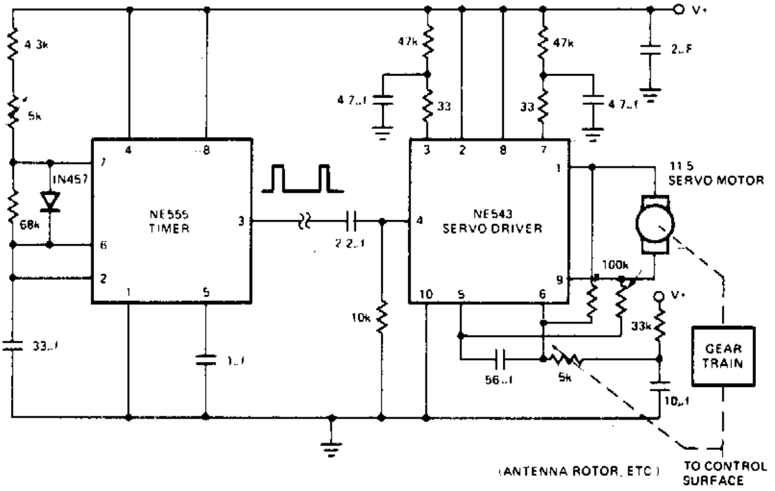


7
6
Out
8
To Output of Board
4
3
In
5
To Input of Board
C1
R2
1
Q1
K1
C2
D1
R1
2
13.6 Vdc

C1 – 3.0 pF Dip Mica
C2 – 0.001 µF Disc
R1 – 1.0 kΩ
R2 – 470 kΩ
D1 – 1N4001
Q1 – 2N4401
K1 – NF2–12 V (Arrow-M) or Equivalent

Fig. 75-2

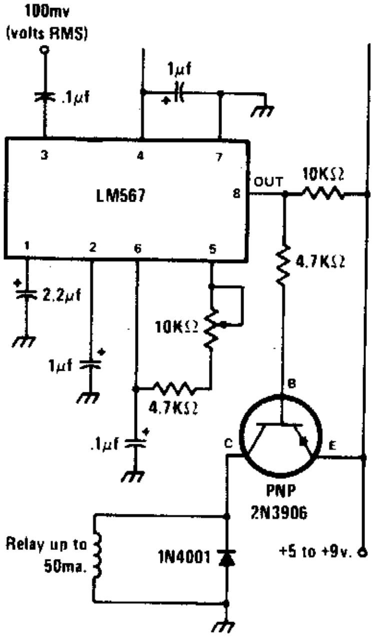
REMOTE CONTROL SERVO SYSTEM



V+
4.3k
5k
1N457
68k
33µF
NE555 TIMER
1µF
2.2µF
10k
47k
4.7µF
33
47k
33
4.7µF
2µF
NE543 SERVO DRIVER
115 SERVO MOTOR
100k
V+
33k
56µF
5k
10µF
GEAR TRAIN
(ANTENNA ROTOR, ETC.)
TO CONTROL SURFACE

Fig. 75-3

TONE-ACTUATED RELAY


100mv
(volts RMS)
.1μf
1μf
LM567
3
4
7
8
1
2
6
5
OUT
10KΩ
4.7KΩ
2.2μf
1μf
10KΩ
4.7KΩ
.1μf
B
C
E
PNP
2N3906
Relay up to
50ma.
1N4001
+5 to +9v.

Circuit Notes

The circuit is built around the LM567 tone decoder IC that requires about 100 millivolts at its operating frequency. The frequency is set by a 10 K variable resistor and can be between 700 and 1500 Hz. When a tone at the set frequency is present, the 567's output goes low to energize a relay through a 2N3906 PNP transistor.

Fig. 75-4

RADIO CONTROL MOTOR SPEED CONTROLLER

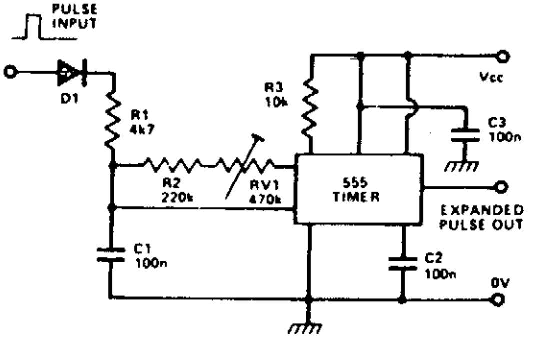

PULSE
INPUT
D1
R1
4k7
R2
220k
RV1
470k
R3
10k
555
TIMER
C3
100n
Vcc
EXPANDED
PULSE OUT
C1
100n
C2
100n
0V

Fig. 75-5

REMOTE ON-OFF SWITCH

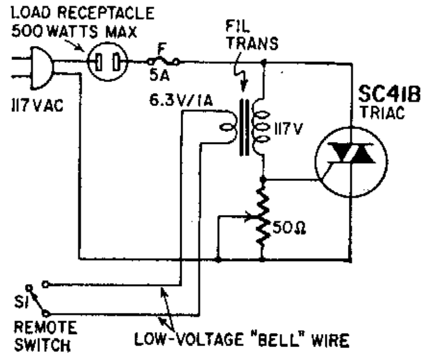
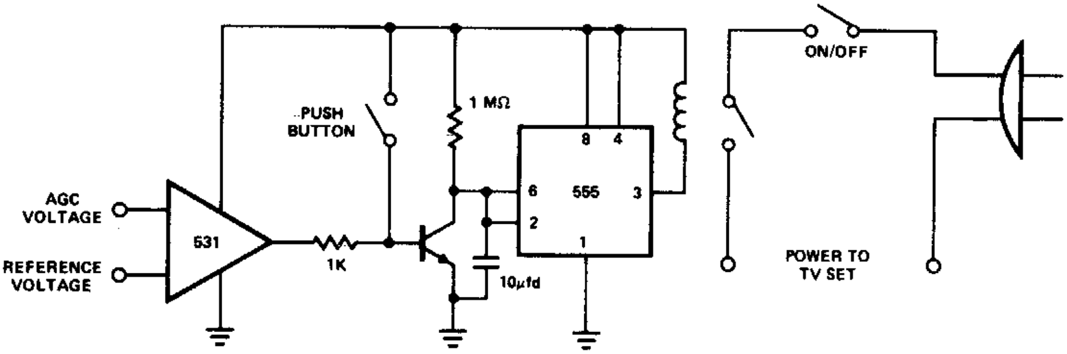

LOAD RECEPTACLE
500 WATTS MAX
F
5A
FIL
TRANS
117 VAC
6.3V/1A
117V
SC41B
TRIAC
50Ω
S1
REMOTE
SWITCH
LOW-VOLTAGE "BELL" WIRE

Fig. 75-6

Circuit Notes

This circuit provides power control without running line-voltage switch leads. The primary of a 6-volt filament transformer is connected between the gate and one of the main terminals of a triac. The secondary is connected to the remote switch through ordinary low-voltage line. With switch open, transformer blocks gate current, prevents the triac from firing and applying power to the equipment. Closing the switch short-circuits the secondary, causing the transformer to saturate and trigger the triac.

AUTOMATIC TURN OFF FOR TV SET


AGC
VOLTAGE
REFERENCE
VOLTAGE
531
1K
PUSH
BUTTON
1 MΩ
10μfd
8
4
6
555
3
2
1
ON/OFF
POWER TO
TV SET

SET TURNS OFF SHORTLY AFTER TV STATION STOPS BROADCASTING

Fig. 75-7

76

Safety and Security Circuits

The sources of the following circuits are contained in the Sources section beginning on page 730. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

Tarry Light
Ground Tester
Ground-Fault Interrupter
Single Source Emergency Lighting System
Power Failure Alarm
Ac Hot Wire Probe
Power Failure Detector
Power-Failure Alarm
Electronic Combination Lock

TARRY LIGHT

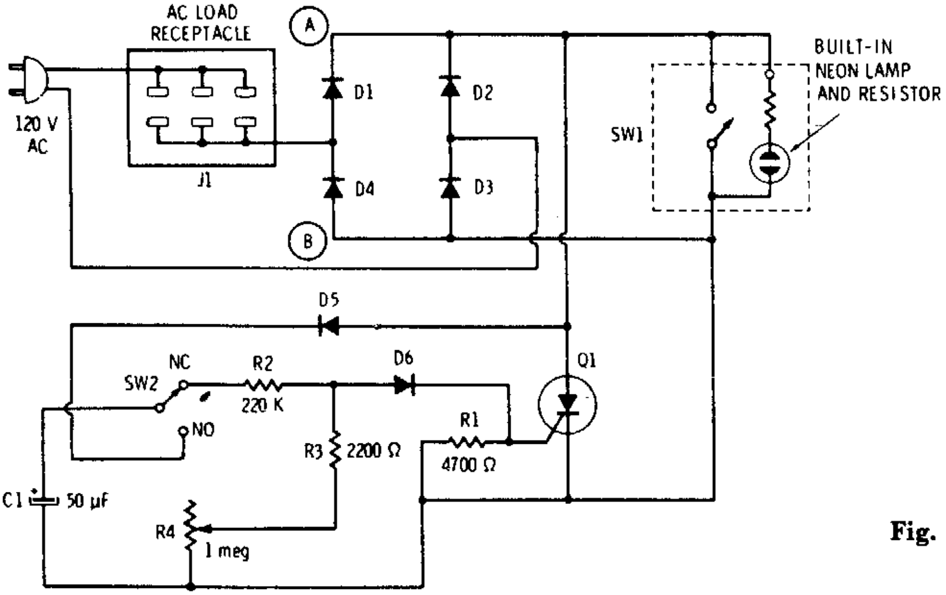
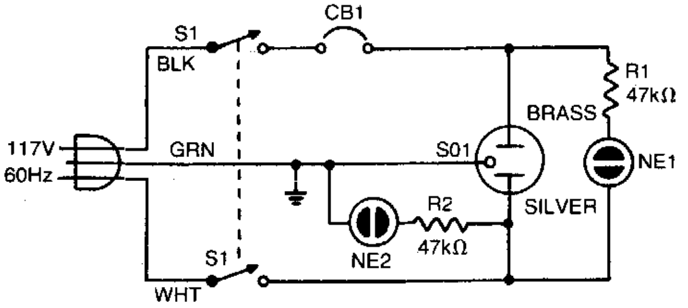

AC LOAD
RECEPTACLE
A
B
D1
D2
D3
D4
BUILT-IN
NEON LAMP
AND RESISTOR
SW1
120 V
AC
J1
D5
NC
R2
220 K
D6
Q1
SW2
NO
R1
4700 Ω
R3
2200 Ω
C1
50 µF
R4
1 meg

Fig. 76-1

Circuit Notes

The push button and potentiometer initiate a time delay that turns a light on then automatically turns it off again after a predetermined time. The potentiometer can be set for a delay of a few seconds to just under three minutes. When the push-button switch SW2 is pressed, capacitor C1 gets charged through D5 to the full dc voltage developed by the diode bridge. When the button is released, the charged capacitor is connected across the series combination of R2, R3, and potentiometer R4 whose setting determines the total resistance and thereby sets the time it takes for the capacitor to discharge. A steering diode, D6, connected to the junction of R2 and R3, and potentiometer R4 whose setting determines The total resistance and thereby sets the time it takes for the capacitor to discharge. Diode, D6 picks off a portion of this decaying dc voltage and applies it to the gates terminal of Q1, the SCR, triggering it into a conductive state. This SCR will remain on as long as there is sufficient voltage on its gate. As soon as this voltage decays below the minimum holding voltage of the SCR, it will turn off on the next line alternation.

GROUND TESTER


CB1
S1
BLK
R1
47kΩ
BRASS
117V
60Hz
GRN
S01
NE1
R2
SILVER
NE2
47kΩ
S1
WHT

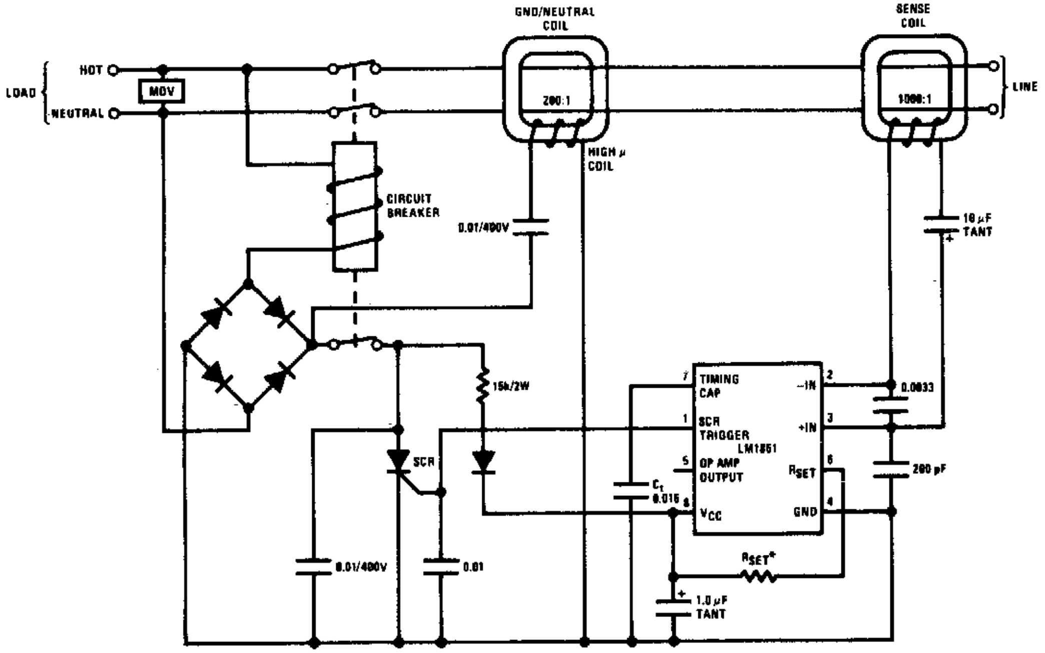
Parts list
S2—DPDT Switch
CB1—10A fuse or circuit breaker
S01—Radio Shack 61-2760, 3 terminal socket
R1, R2, R3—47kΩ, ½W
NE1, NE2, NE3—GE NE-2
S1—SPDT, Lafayette 34P0238V

Fig. 76-2

Circuit Notes

This circuit checks the reliability of appliances so that the equipment may be used safely. The test circuit must be plugged into a properly wired three terminal wall outlet. When a two-lead or three-lead appliance is plugged into circuit outlet S01, neon lamps NE1 and NE2 will light if the appliance is safe. If neon NE2 is lit the appliance is dangerous, because the neutral lead is 110 Vac above ground.

GROUND-FAULT INTERRUPTER (120 Hz NEUTRAL TRANSFORMER APPROACH)


GND/NEUTRAL COIL
SENSE COIL
LOAD
HOT
NEUTRAL
MOV
200:1
1000:1
LINE
HIGH μ COIL
CIRCUIT BREAKER
0.01/400V
10 μF TANT
15k/2W
SCR
7 TIMING CAP
1 SCR TRIGGER
LM1851
5 OP AMP OUTPUT
8 VCC
2 −IN
3 +IN
6 RSET
4 GND
0.0033
200 pF
Ct 0.015
RSET*
1.0 μF TANT
0.01/400V
0.01

* Adjust R_{SET} for desired sensitivity

Fig. 76-3

SINGLE SOURCE EMERGENCY LIGHTING SYSTEM

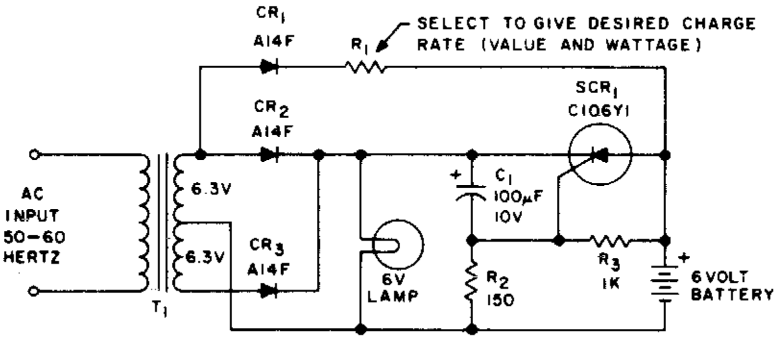
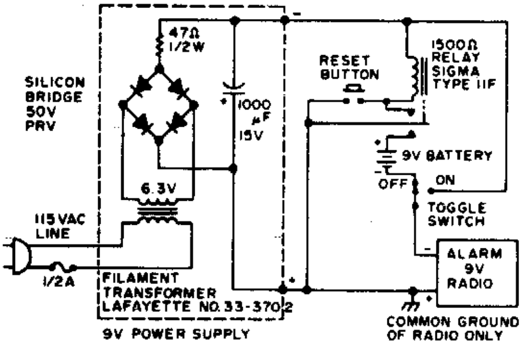

CR1
A14F
R1
SELECT TO GIVE DESIRED CHARGE RATE (VALUE AND WATTAGE)
SCR1
C106Y1
CR2
A14F
AC INPUT 50–60 HERTZ
6.3V
6.3V
CR3
A14F
T1
C1
100μF
10V
6V LAMP
R2
150
R3
1K
6 VOLT BATTERY

Fig. 76-4

Circuit Notes

This emergency lighting system maintains a 6 volt battery at full charge and switches automatically from the ac supply to the battery.

POWER FAILURE ALARM

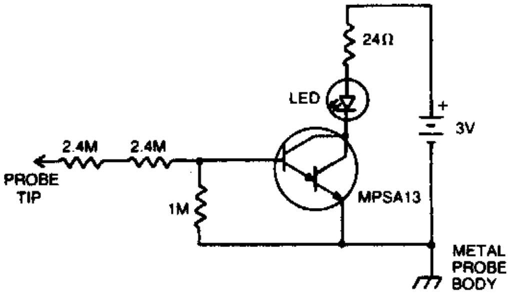

47Ω
1/2W
SILICON BRIDGE 50V PRV
1000 μF 15V
6.3V
115VAC LINE
1/2A
FILAMENT TRANSFORMER LAFAYETTE NO. 33-3702
9V POWER SUPPLY
RESET BUTTON
1500Ω RELAY SIGMA TYPE 11F
9V BATTERY
OFF
ON
TOGGLE SWITCH
ALARM 9V RADIO
COMMON GROUND OF RADIO ONLY

Circuit Notes

If the power fails, the radio alarm goes on. No loud siren, bell, or whistle. Even if the power is restored, the alarm stays on until RESET button is pushed.

Fig. 76-5

AC HOT WIRE PROBE


24Ω
LED
3V
2.4M
2.4M
PROBE TIP
1M
MPSA13
METAL PROBE BODY

Circuit Notes

Insert the probe tip into either terminal of an ac outlet and hold the probe body against anything that the circuit ground is connected to. The LED will glow when the hot terminal is touched. Two 2.4 M resistors are used in the probe tip for safety (redundancy) reasons.

Fig. 76-6

POWER FAILURE DETECTOR

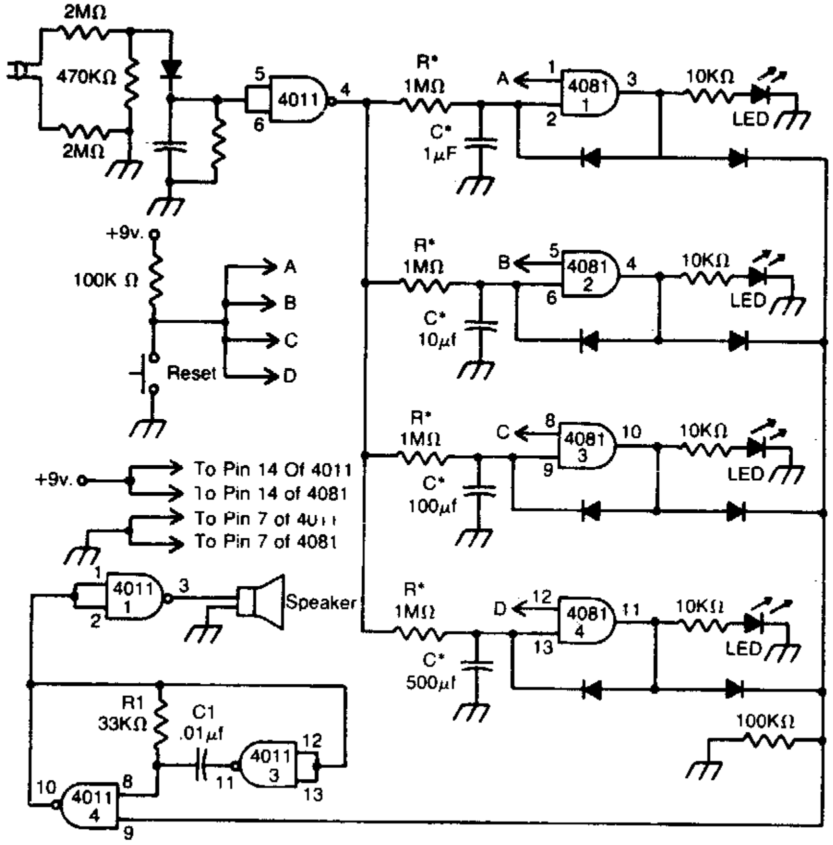

2MΩ
470KΩ
2MΩ
5
6
4011
4
R*
1MΩ
C*
1μF
A
1
2
4081
1
3
10KΩ
LED
+9v.
100K Ω
A
B
C
D
Reset
R*
1MΩ
C*
10μf
B
5
6
4081
2
4
10KΩ
LED
R*
1MΩ
C*
100μf
C
8
9
4081
3
10
10KΩ
LED
+9v.
To Pin 14 Of 4011
To Pin 14 of 4081
To Pin 7 of 4011
To Pin 7 of 4081
1
2
4011
1
3
Speaker
R*
1MΩ
C*
500μf
D
12
13
4081
4
11
10KΩ
LED
100KΩ
R1
33KΩ
C1
.01μf
4011
3
12
13
11
10
8
4011
4
9

Fig. 76-7

Circuit Notes

This circuit indicates that a power outage occured for 1, 10, 100, and 500 seconds with the values given for R* and C*. After a power failure, the circuit can be reset by pushing the Reset button.

POWER FAILURE ALARM

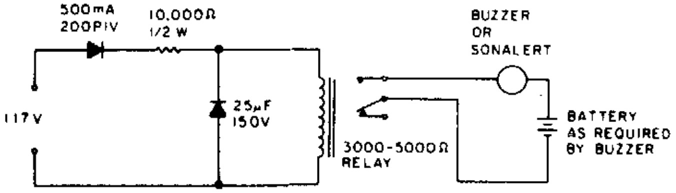
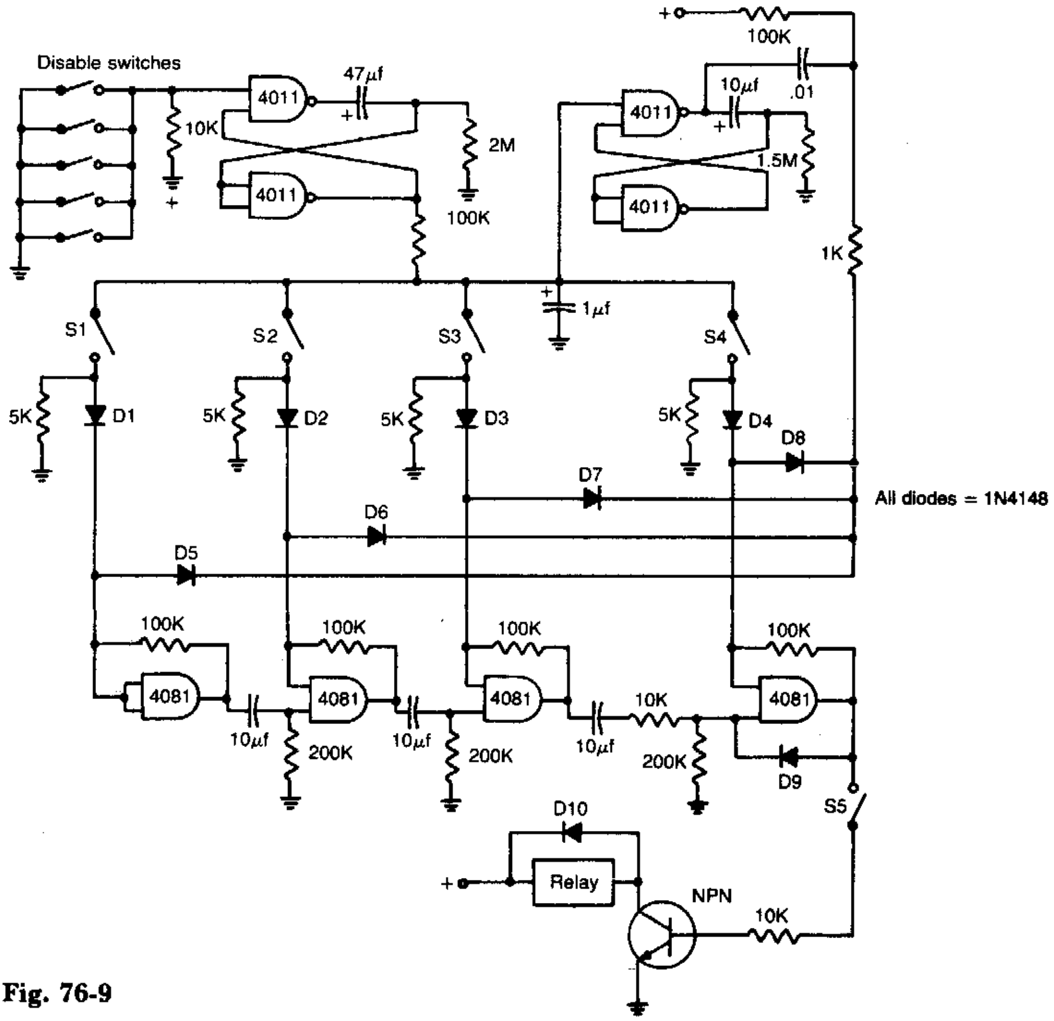

500mA
200PIV
10,000Ω
1/2 W
117V
25μF
150V
3000-5000Ω
RELAY
BUZZER
OR
SONALERT
BATTERY
AS REQUIRED
BY BUZZER

Fig. 76-8

Circuit Notes

While the power is on, the relay is held open, but when the power fails the buzzer-circuit contacts close.

ELECTRONIC COMBINATION LOCK



Disable switches
10K
+
4011
4011
47μf
+
2M
100K
100K
+
4011
4011
10μf
+
.01
1.5M
1K
+
1μf
S1
S2
S3
S4
5K
D1
5K
D2
5K
D3
5K
D4
D8
D7
D6
D5
All diodes = 1N4148
100K
100K
100K
100K
4081
4081
4081
4081
10μf
200K
10μf
200K
10μf
10K
200K
D9
S5
D10
+
Relay
NPN
10K

Fig. 76-9

Circuit Notes

Switches S1 through S5 must be operated in rapid sequence to operate the lock. They can be any numbers on a 10-button switch pad. If an incorrect button is pushed, alarm sounds and the circuit is disabled for two minutes.

77

Sample and Hold Circuits

The sources of the following circuits are contained in the Sources section beginning on page 730. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

Peak Detect and Hold
Low Drift Sample and Hold
JFET Sample and Hold
High Speed Sample and Hold Amplifier
High Speed Sample and Hold
High Speed Sample and Hold
Sample and Hold with Offset Adjustment
Differential Hold
× 1000 Sample and Hold
Sample and Hold
High Accuracy Sample and Hold
High Speed Sample and Hold

PEAK DETECT AND HOLD

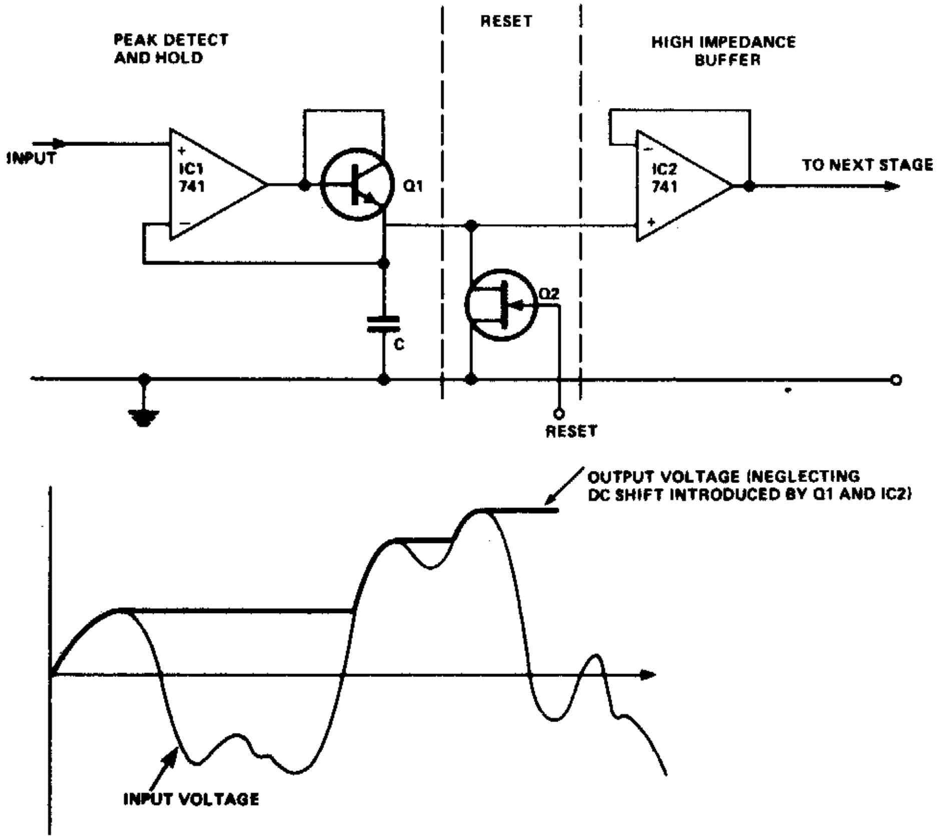
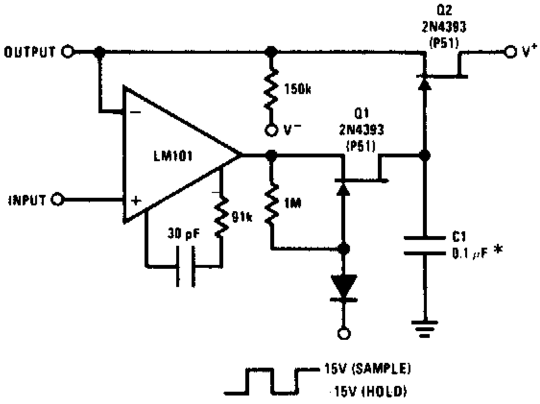

PEAK DETECT AND HOLD
RESET
HIGH IMPEDANCE BUFFER
INPUT
+
IC1
741
−
Q1
C
Q2
−
IC2
741
+
TO NEXT STAGE
RESET
OUTPUT VOLTAGE (NEGLECTING DC SHIFT INTRODUCED BY Q1 AND IC2)
INPUT VOLTAGE

Fig. 77-1

Circuit Notes

If the voltage at the input exceeds the voltage on the capacitor, then the output of the 741 goes positive, the diode conducts, and the capacitor is charged up to the input voltage-forward voltage drop of diode. When the voltage at the input is less than that on the capacitor, the output of the 741 goes negative, and the diode cuts off. To prevent the capacitor from discharging through the input resistance of the next stage, a high input impedance buffer stage (IC2) is used. The circuit can be reset by means of a FET or similar high impedance device connected across the capacitor.

LOW DRIFT SAMPLE AND HOLD


Q2
2N4393
(P51)
OUTPUT
V+
150k
V−
Q1
2N4393
(P51)
LM101
INPUT
1M
91k
30 pF
C1
0.1 μF *
15V (SAMPLE)
15V (HOLD)

*Polycarbonate dielectric capacitor

Circuit Notes

The JFETs, Q1 and Q2, provide complete buffering to C1, the sample and hold capacitor. During sample, Q1 is turned on and provides a path, $r_{ds(on)}$, for charging C1. During hold, Q1 is turned off, thus leaving Q1 $I_{D(off)}$ (< 100 pA) and Q2 I_{GSS} (< 100 pA) as the only discharge paths. Q2 serves a buffering function so feedback to the LM101 and output current are supplied from its source.

Fig. 77-2

JFET SAMPLE AND HOLD

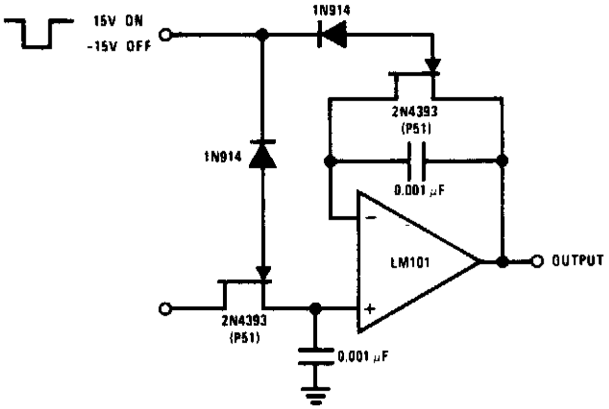
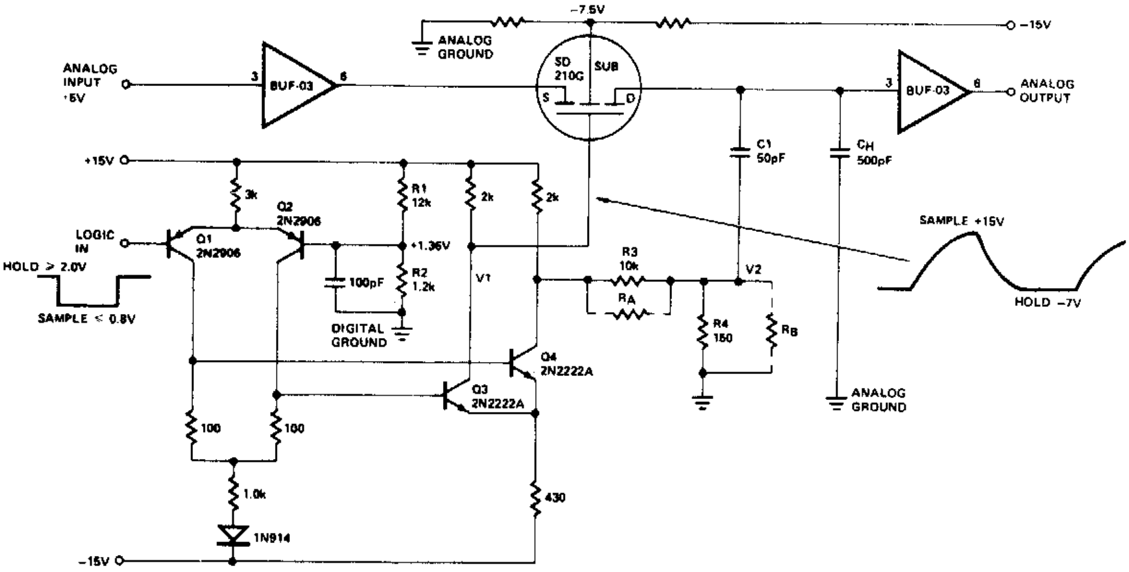

15V ON
-15V OFF
1N914
1N914
2N4393
(P51)
0.001 μF
LM101
OUTPUT
2N4393
(P51)
0.001 μF

Fig. 77-3

Circuit Notes

The logic voltage is applied simultaneously to the sample and hold JFETs. By matching input impedance and feedback resistance and capacitance, errors due to $r_{ds(on)}$ of the JFETs are minimized.

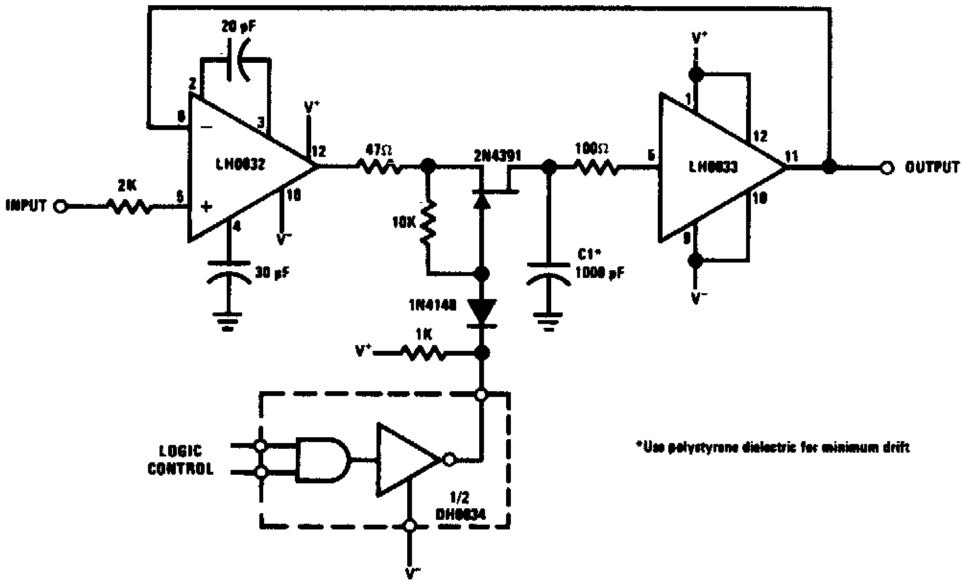
HIGH SPEED SAMPLE AND HOLD AMPLIFIER



-7.5V
-15V
ANALOG GROUND
ANALOG INPUT ±5V
3
BUF-03
6
SD 210G
SUB
S
D
3
BUF-03
6
ANALOG OUTPUT
+15V
3k
Q2 2N2906
R1 12k
2k
2k
C1 50pF
CH 500pF
LOGIC IN
Q1 2N2906
+1.36V
HOLD ≥ 2.0V
SAMPLE ≤ 0.8V
100pF
R2 1.2k
V1
R3 10k
V2
RA
SAMPLE +15V
HOLD -7V
DIGITAL GROUND
R4 150
RB
Q4 2N2222A
Q3 2N2222A
ANALOG GROUND
100
100
1.0k
430
1N914
-15V

Fig. 77-4

HIGH SPEED SAMPLE AND HOLD



20 pF
2
6
3
V+
12
LH0032
47Ω
2N4391
100Ω
5
V+
1
12
LH0033
11
OUTPUT
INPUT
2K
5
10
V-
4
10K
30 pF
C1* 1000 pF
9
10
V-
1N4148
1K
V+
LOGIC CONTROL
1/2 DH0034
V-

*Use polystyrene dielectric for minimum drift

Fig. 77-5

HIGH SPEED SAMPLE AND HOLD

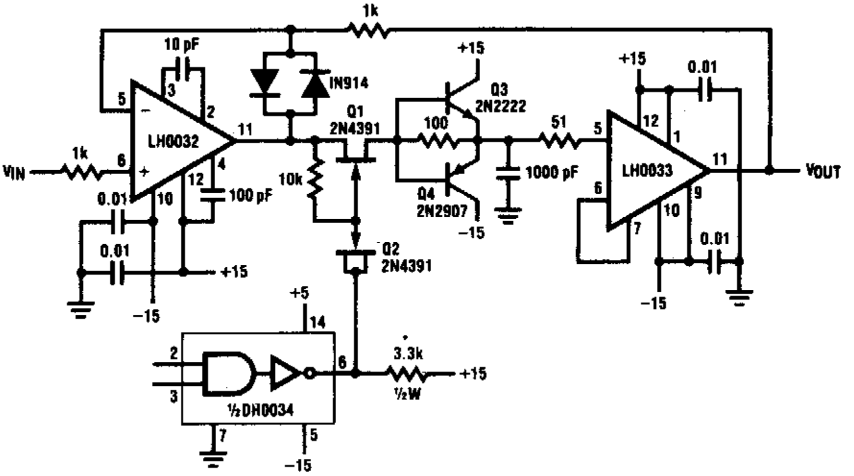
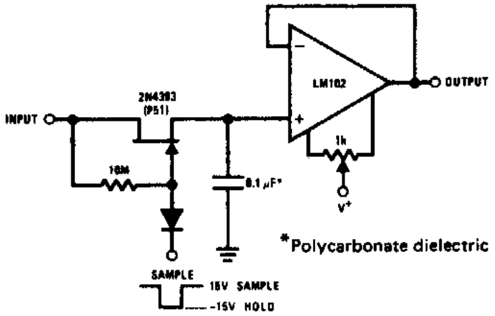

1k
10 pF
+15
IN914
Q3
2N2222
+15
0.01
5
3
2
Q1
2N4391
100
51
5
12
1
LH0032
11
11
VIN
1k
6
4
10k
1000 pF
LH0033
VOUT
0.01
12
100 pF
Q4
2N2907
6
9
10
10
-15
7
0.01
Q2
2N4391
0.01
+15
-15
-15
+5
14
3.3k
2
6
+15
½W
3
½DH0034
7
5
-15

Fig. 77-6

Circuit Notes

This circuit exhibits a 10 V acquisition time of 900 ns to 0.1% accuracy and a droop rate of only 100 μV/ms at 25° C ambient condition. An even faster acquisition time can be obtained using a smaller value hold-capacitor. By decreasing the value from 1000 pF to 220 pF, the acquisition time improves to 500 ns for a 10 V step. However, the droop rate increases to 500 μV/ms.

SAMPLE AND HOLD WITH OFFSET ADJUSTMENT


LM102
OUTPUT
2N4393
(P51)
INPUT
1k
10M
0.1 μF*
V⁺
SAMPLE
15V SAMPLE
-15V HOLD

*Polycarbonate dielectric

Fig. 77-7

Circuit Notes

The 2N4393 JFET was selected because of its low I_{GSS} (< 100 pA), very low $I_{D(off)}$ (< 100 pA) and low pinchoff voltage. Leakages of this level put the burden of circuit performance on clean, solder-resin free, low leakage circuit layout.

DIFFERENTIAL HOLD

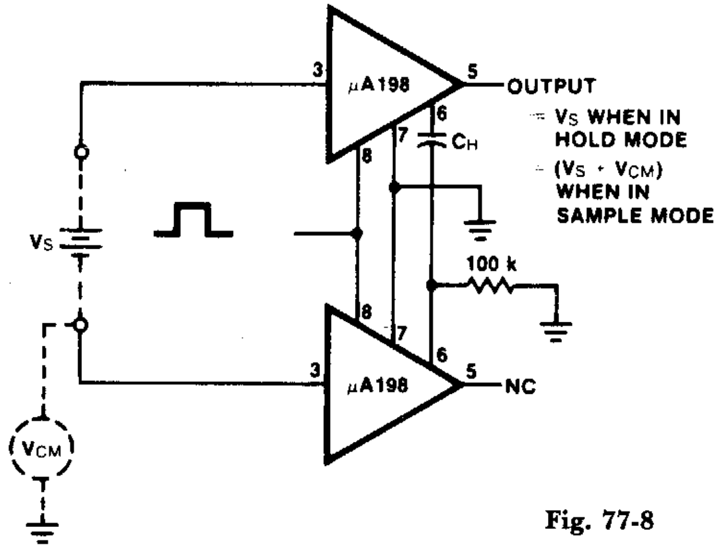
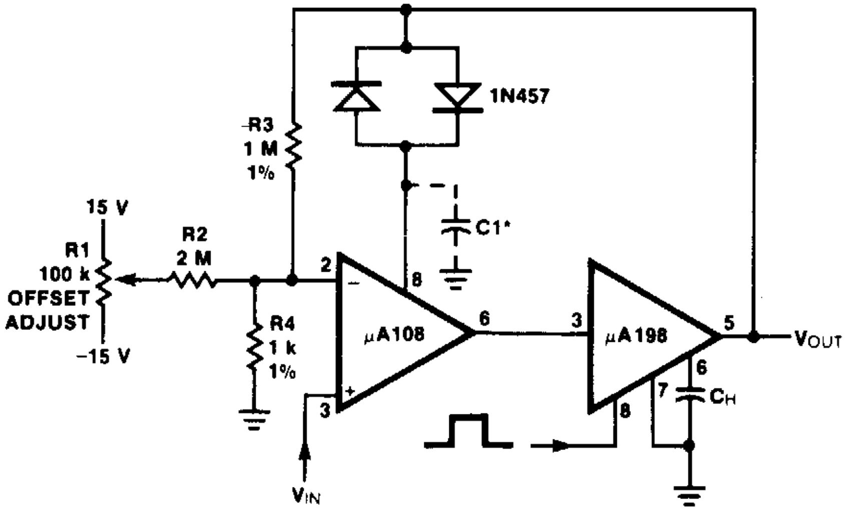

3
μA198
5
OUTPUT
≈ Vs WHEN IN HOLD MODE
≈ (Vs + VCM) WHEN IN SAMPLE MODE
6
7
CH
8
Vs
100 k
8
7
6
3
μA198
5
NC
VCM

Fig. 77-8

× 1000 SAMPLE AND HOLD


1N457
R3
1 M
1%
C1*
15 V
R1
100 k
OFFSET
ADJUST
−15 V
R2
2 M
R4
1 k
1%
2
−
+
3
8
μA108
6
3
μA198
5
VOUT
6
7
CH
8
VIN

Notes

For lower gains, the μA108 must be frequency compensated

Use $\approx \frac{100}{A_V}$ pF from comp 2 to ground

Fig. 77-9

SAMPLE AND HOLD

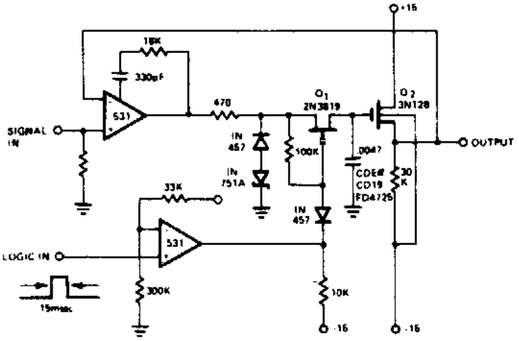
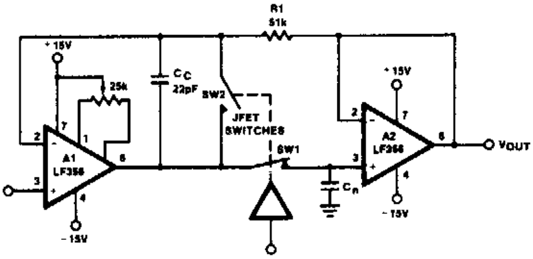

+15
10K
330pF
470
Q1
2N3819
Q2
3N128
SIGNAL IN
531
IN 457
IN 751A
100K
.0047
CDE# CD19 FD4725
30K
OUTPUT
33K
LOGIC IN
531
15mmax
300K
IN 457
10K
-15
-15

Fig. 77-10

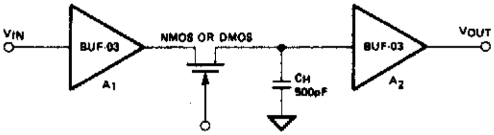
HIGH ACCURACY SAMPLE AND HOLD


R1
51k
+15V
25k
CC
22pF
SW2
JFET SWITCHES
SW1
A1
LF356
A2
LF356
+15V
-15V
-15V
CH
VOUT

- By closing the loop through A2 the V_{OUT} accuracy will be determined uniquely by A1. No V_{OS} adjust required for A2.
- T_A can be estimated by same considerations as previously but, because of the added on propagation delay in the feedback loop (A2) the overshoot is not negligible.
- Overall system slower than fast sample and hold.
- R1, C_C: additional compensation
- Use LF356 for
  - Δ Fast settling time
  - Δ Low V_{OS}

Fig. 77-11

HIGH SPEED SAMPLE AND HOLD


VIN
BUF-03
A1
NMOS OR DMOS
CH
500pF
BUF-03
A2
VOUT

ICHARGE OF BUF-03 IS ±60mA. THEREFORE THE SLEW RATE INTO A 500pF HOLD CAPACITOR WILL BE 120V/μSEC. THUS THE SLEW RATE OF THE SAMPLE AND HOLD CIRCUIT IS LIMITED BY THE CAPACITOR CHARGING TIME.

Fig. 77-12

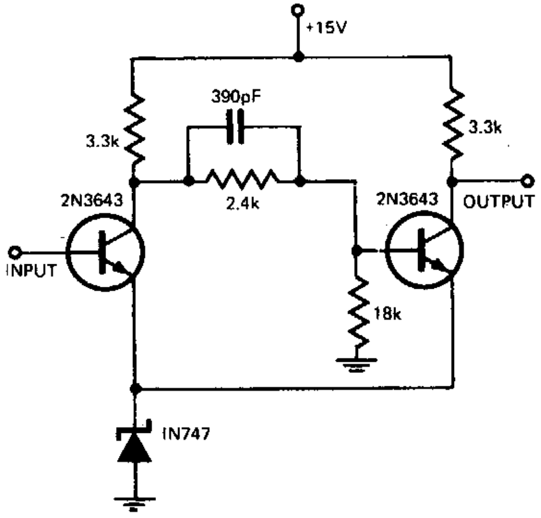
78

Schmitt Triggers

The sources of the following circuits are contained in the Sources section beginning on page 730. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

Schmitt Trigger Without Hysteresis
Schmitt Trigger with Programmable Hysteresis
Schmitt Trigger (Zero Crossing Detector with Hysteresis)
Schmitt Trigger

SCHMITT TRIGGER WITHOUT HYSTERESIS


+15V
390pF
3.3k
3.3k
2.4k
2N3643
2N3643
OUTPUT
INPUT
18k
IN747

Circuit Notes

By replacing the common-emitter resistor in a conventional Schmitt by a zener diode, the hysteresis normally associated with these circuits is eliminated.

Fig. 78-1

SCHMITT TRIGGER WITH PROGRAMMABLE HYSTERESIS

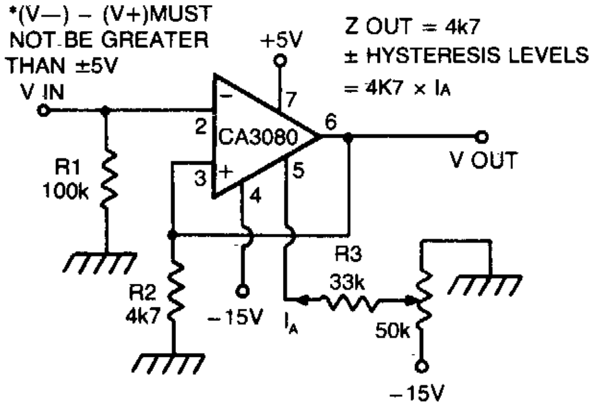
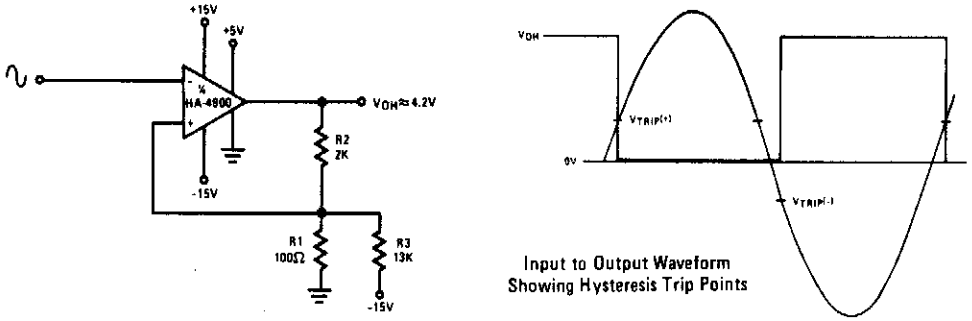

*(V—) - (V+)MUST
NOT BE GREATER
THAN ±5V
V IN
+5V
Z OUT = 4k7
± HYSTERESIS LEVELS
= 4K7 × IA
7
2
CA3080
6
3
5
4
V OUT
R1
100k
R2
4k7
-15V
IA
R3
33k
50k
-15V

Fig. 78-2

Circuit Notes

CA 3088 is used as a versatile Schmitt trigger. The size of the hysteresis levels is determined by I_A that flows out of the amplifier's output and through R2. Increasing I_A increases hysteresis and vice versa. The positive and negative hysteresis levels are symmetrical about 0 V.

SCHMITT TRIGGER (ZERO CROSSING DETECTOR WITH HYSTERESIS)


+15V
+5V
¼
HA-4900
-15V
$V_{OH} \approx 4.2V$
R2
2K
R1
100Ω
R3
13K
-15V
V_{OH}
0V
$V_{TRIP(+)}$
$V_{TRIP(-)}$

Input to Output Waveform
Showing Hysteresis Trip Points

Fig. 78-3

Circuit Notes

This circuit has a 100 mV hysteresis which can be used in applications where very fast transition times are required at the output even though the signal is very slow. The hysteresis loop also reduces false triggering due to noise on the input. The waveforms show the trip points developed by the hysteresis loop.

SCHMITT TRIGGER

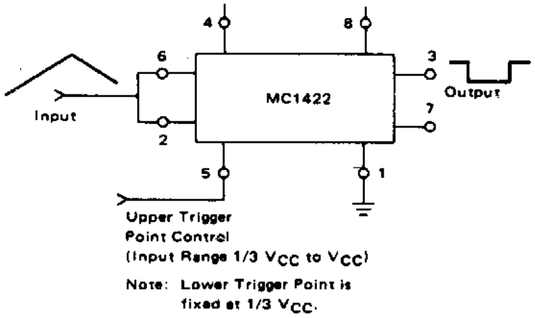

4
8
6
3
MC1422
Output
7
Input
2
5
1
Upper Trigger
Point Control
(Input Range 1/3 V_{CC} to V_{CC})
Note: Lower Trigger Point is
fixed at 1/3 V_{CC}.

Fig. 78-4

Circuit Notes

The lower trigger point is fixed at ⅓ Vcc, but the upper trigger point is adjustable by means of Pin 5 from ⅓ Vcc to slightly less than Vcc. The Schmitt trigger will operate with input frequencies up to 50 kHz.

79

Smoke and Flame Detectors

The sources of the following circuits are contained in the Sources section beginning on page 730. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

Photoelectric Smoke Detector (Non-Latching)
1.9 V Battery Operated Ionization Type Smoke Detector
Line-Operated Photo-Electric Smoke Alarm Using Light Sensitive Resistor
(Includes Detection of Open-Circuited LED)

PHOTOELECTRIC SMOKE DETECTOR (NON-LATCHING)

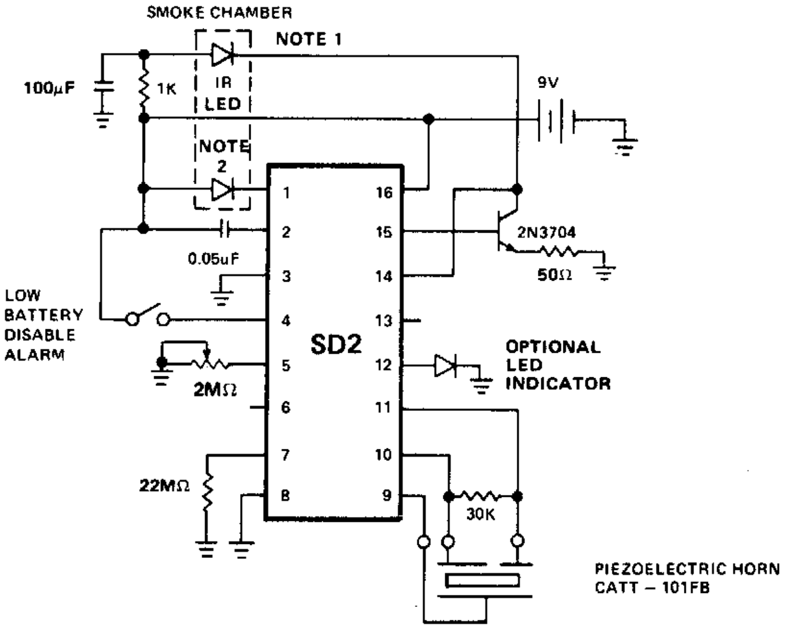

SMOKE CHAMBER
NOTE 1
100μF
1K
IR LED
9V
NOTE 2
1
2
3
4
5
6
7
8
16
15
14
13
12
11
10
9
SD2
2N3704
0.05uF
50Ω
LOW BATTERY DISABLE ALARM
OPTIONAL LED INDICATOR
2MΩ
22MΩ
30K
PIEZOELECTRIC HORN CATT – 101FB

Fig. 79-1

Notes: 1. IR Diode RCA Type SG 1010A or Spectronics Type SE 5455-4
Clairex Type CLED-1
2. IR Photo detectors Vactec VTS4085

Circuit Notes

The LED predriver output pulses an external transistor which in turn, switches on the infrared light emitting diode at a very low duty cycle. The desired IR LED pulse period is determined by the value of the external timing resistor. The Smoke Sensitivity is adjustable through a trimmer resistor which varies the IR LED pulse width. The light sensing element is a silicon photovoltaic cell which is held at near zero bias to minimize leakage currents. The circuit can detect signals as low as 1 mV and generate an alarm. The IR LED pulse repetition rate increases when smoke is detected.

1.9 V BATTERY OPERATED IONIZATION TYPE SMOKE DETECTOR

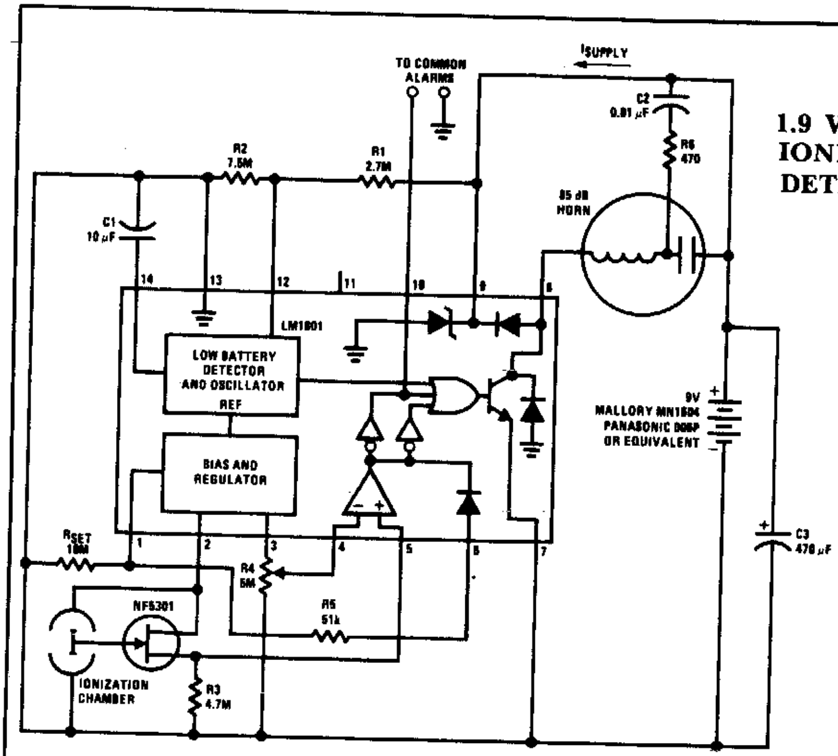

TO COMMON ALARMS
I_{SUPPLY}
C2
0.01 μF
R6
470
R2
7.5M
R1
2.7M
85 dB
HORN
C1
10 μF
14
13
12
11
10
9
8
LM1801
LOW BATTERY DETECTOR AND OSCILLATOR
REF
BIAS AND REGULATOR
9V
MALLORY MN1604
PANASONIC 006P
OR EQUIVALENT
R_{SET}
10M
1
2
3
4
5
6
7
R4
5M
R5
51k
NF5301
IONIZATION CHAMBER
R3
4.7M
C3
470 μF

Fig. 79-2

LINE-OPERATED PHOTO ELECTRIC SMOKE ALARM USING LIGHT SENSITIVE RESISTOR (INCLUDES DETECTION OF OPEN-CIRCUITED LED)

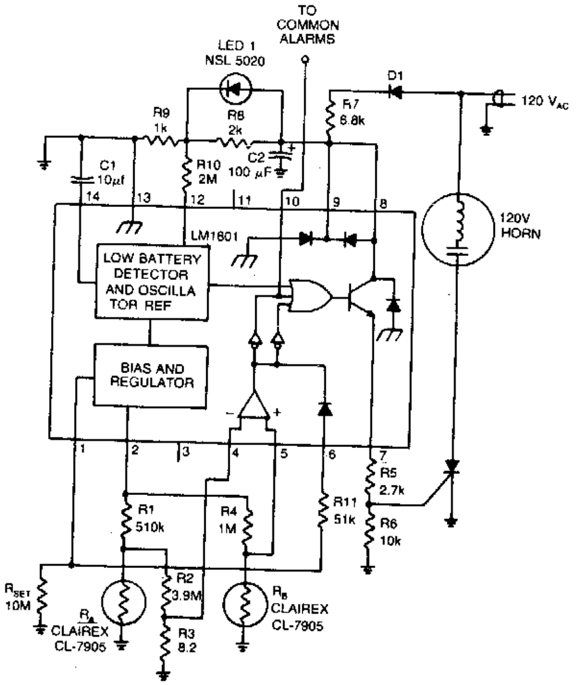

TO COMMON ALARMS
LED 1
NSL 5020
D1
120 V_{AC}
R7
6.8k
R9
1k
R8
2k
C2
100 μF
C1
10μf
R10
2M
14
13
12
11
10
9
8
LM1601
LOW BATTERY DETECTOR AND OSCILLA TOR REF
BIAS AND REGULATOR
120V
HORN
1
2
3
4
5
6
7
R5
2.7k
R1
510k
R4
1M
R11
51k
R6
10k
R_{SET}
10M
R_A
CLAIREX
CL-7905
R2
3.9M
R3
8.2
R_B
CLAIREX
CL-7905

Fig. 79-3

1.9 V BATTERY OPERATED IONIZATION TYPE SMOKE DETECTOR

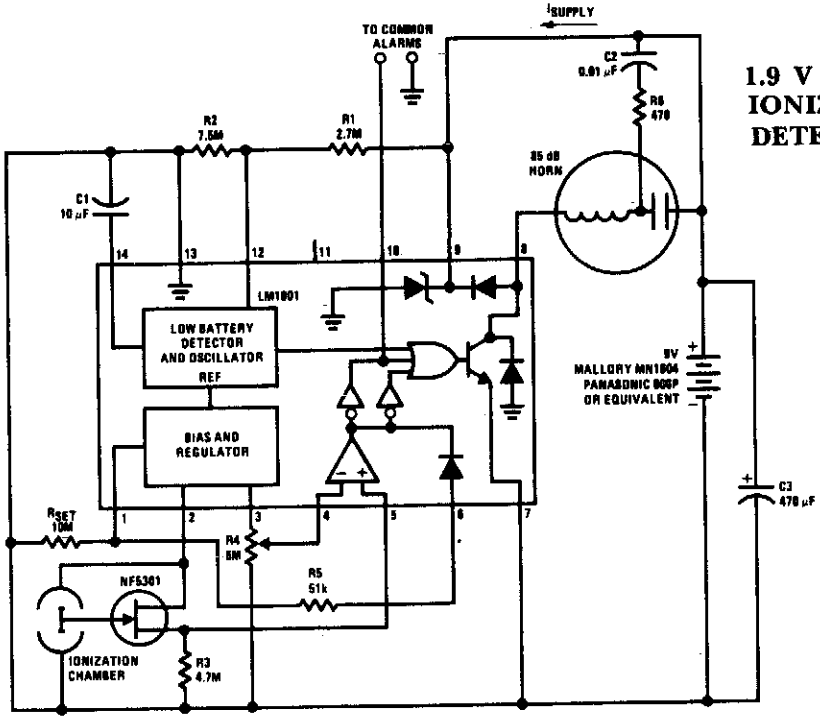

I_SUPPLY
TO COMMON ALARMS
C2 0.01 µF
R6 470
85 dB HORN
R2 7.5M
R1 2.7M
C1 10 µF
14
13
12
11
10
9
8
LM1801
LOW BATTERY DETECTOR AND OSCILLATOR
REF
BIAS AND REGULATOR
9V MALLORY MN1604 PANASONIC 006P OR EQUIVALENT
C3 470 µF
R_SET 10M
1
2
3
4
5
6
7
R4 5M
R5 51k
NF5301
IONIZATION CHAMBER
R3 4.7M

Fig. 79-2

LINE-OPERATED PHOTO ELECTRIC SMOKE ALARM USING LIGHT SENSITIVE RESISTOR (INCLUDES DETECTION OF OPEN-CIRCUITED LED)

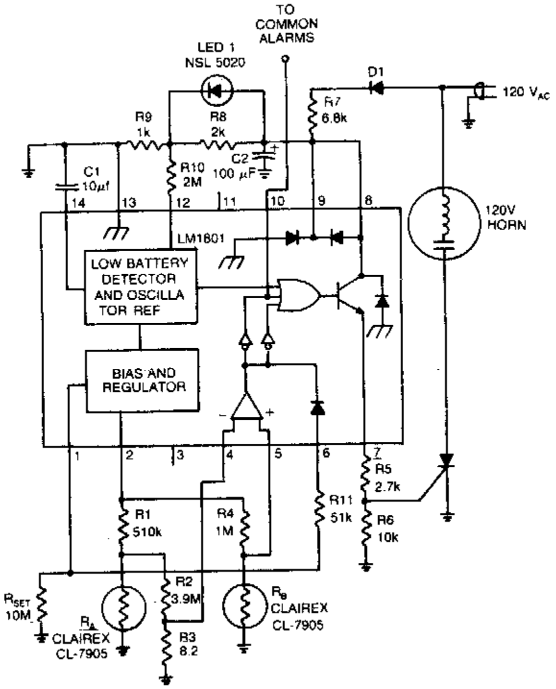

TO COMMON ALARMS
LED 1 NSL 5020
D1
120 V_AC
R7 6.8k
R9 1k
R8 2k
C2 100 µF
C1 10µf
R10 2M
14
13
12
11
10
9
8
LM1801
120V HORN
LOW BATTERY DETECTOR AND OSCILLA TOR REF
BIAS AND REGULATOR
1
2
3
4
5
6
7
R5 2.7k
R6 10k
R11 51k
R1 510k
R4 1M
R2 3.9M
R3 8.2
R_SET 10M
R_A CLAIREX CL-7905
R_B CLAIREX CL-7905

Fig. 79-3

80

Sound Effect Circuits

The sources of the following circuits are contained in the Sources section beginning on page 730. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

- Voltage-Controlled Amplifier or Tremolo Circuit
- Music Synthesizer
- Preprogrammed Single-Chip Microcontroller for Musical Organ
- Musical Envelope Generator and Modulator
- Stereo Reverb System
- Tone Burst Generator
- Musical Chime Generator
- Sound Effect Generator
- Programmable Bird Sounds
- Stereo Reverb Enhancement System
- Siren/Space War/Phasor Gun
- Four Channel Synthesizer

VOLTAGE-CONTROLLED AMPLIFIER OR TREMOLO CIRCUIT

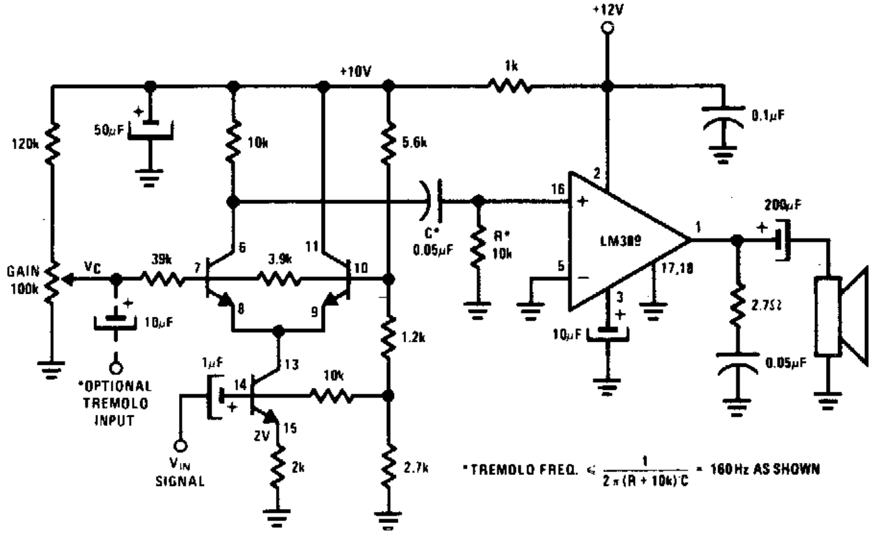

+12V
+10V
1k
0.1μF
120k
50μF
10k
5.6k
16
2
C*
0.05μF
R*
10k
LM389
1
200μF
GAIN
100k
VC
39k
7
6
3.9k
11
10
5
17,18
8
9
10μF
3
2.7Ω
1.2k
10μF
0.05μF
*OPTIONAL
TREMOLO
INPUT
1μF
14
13
10k
15
2V
VIN
SIGNAL
2k
2.7k
*TREMOLO FREQ. ≤ $\frac{1}{2\pi(R+10k)C}$ = 160Hz AS SHOWN

Fig. 80-1

Circuit Notes

The transistors form a differential pair with an active current-source tail. This configuration, known technically as a variable-transconductance multiplier, has an output proportional to the product of the two input signals. Multiplication occurs due to the dependence of the transistor transconductance on the emitter current bias. Tremolo (amplitude modulation of an audio frequency by a sub-audio oscillator—normally 5-15 Hz) applications require feeding the low frequency oscillator signal into the optional input shown. The gain control pot maybe set for optimum depth.

MUSIC SYNTHESIZER

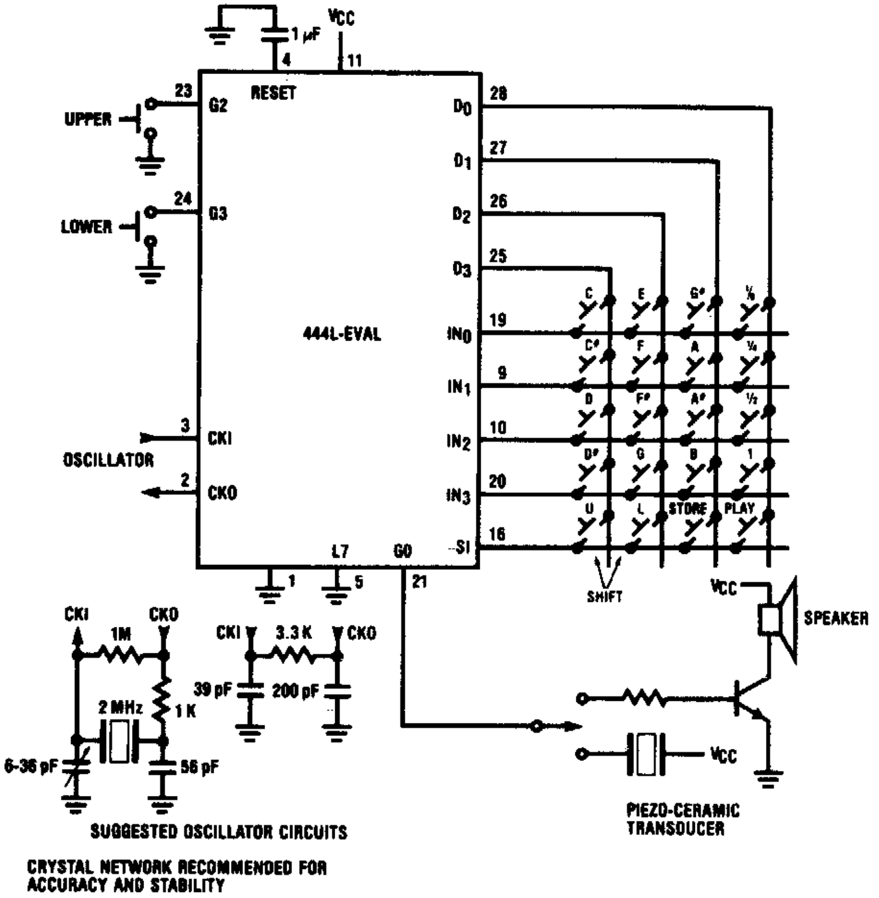

VCC
1 µF
4
11
RESET
23
G2
UPPER
24
G3
LOWER
444L-EVAL
3
CKI
OSCILLATOR
2
CKO
L7
G0
1
5
21
28
D0
27
D1
26
D2
25
D3
19
IN0
9
IN1
10
IN2
20
IN3
16
SI
C
E
G#
⅛
C#
F
A
¼
D
F#
A#
½
D#
G
B
1
U
L
STORE
PLAY
SHIFT
VCC
SPEAKER
VCC
PIEZO-CERAMIC TRANSDUCER
CKI
CKO
1M
2 MHz
1 K
6-36 pF
56 pF
CKI
3.3 K
CKO
39 pF
200 pF
SUGGESTED OSCILLATOR CIRCUITS
CRYSTAL NETWORK RECOMMENDED FOR ACCURACY AND STABILITY

Fig. 80-2

Circuit Notes

Three modes of operation are available in the music synthesizer mode: play a note, play one of four stored tunes, or record a tune for subsequent replay.

PREPROGRAMMED SINGLE-CHIP MICROCONTROLLER FOR MUSICAL ORGAN

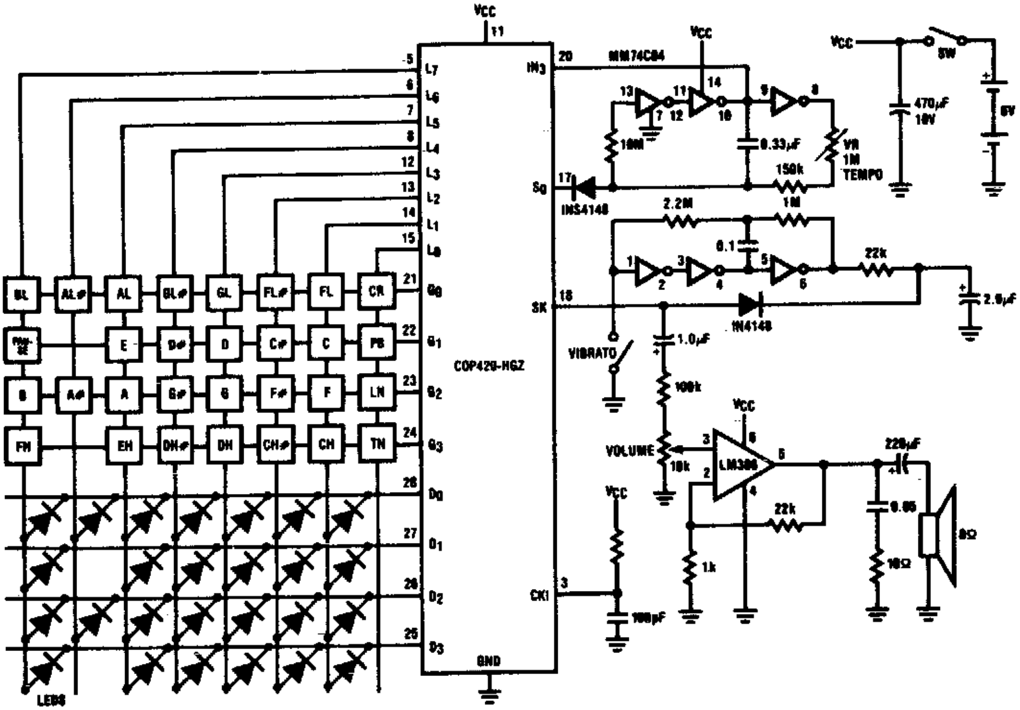

VCC
11
COP420-HGZ
L7
L6
L5
L4
L3
L2
L1
L0
G0
G1
G2
G3
D0
D1
D2
D3
IN3
S0
SK
CKI
GND
5
6
7
8
12
13
14
15
21
22
23
24
28
27
26
25
20
17
18
3
BL
AL#
AL
GL#
GL
FL#
FL
CR
PAU-SE
E
D#
D
C#
C
PB
B
A#
A
G#
G
F#
F
LN
FH
EH
DH#
DH
CH#
CH
TN
LEDS
MM74C04
VCC
13
12
11
10
14
7
9
8
10M
0.33μF
VR
1M
TEMPO
150k
IN54148
2.2M
1M
0.1
1
2
3
4
5
6
22k
2.0μF
IN4148
VIBRATO
1.0μF
100k
VOLUME
10k
VCC
LM386
3
2
6
5
4
22k
1k
220μF
0.05
10Ω
8Ω
VCC
100pF
VCC
SW
470μF
16V
6V

Fig. 80-3

Circuit Notes

Twenty-five musical keys and 25 LEDs are provided to denote F to F″ with half notes in between. Memory can store a played tune. There are ten preprogrammed tunes (each has an average of 55 notes) masked in the chip. Any tune can be recalled by depressing the Tune Button followed by the corresponding Sharp Key. In learn mode, the player can learn the ten preprogrammed tunes.

MUSICAL ENVELOPE GENERATOR AND MODULATOR

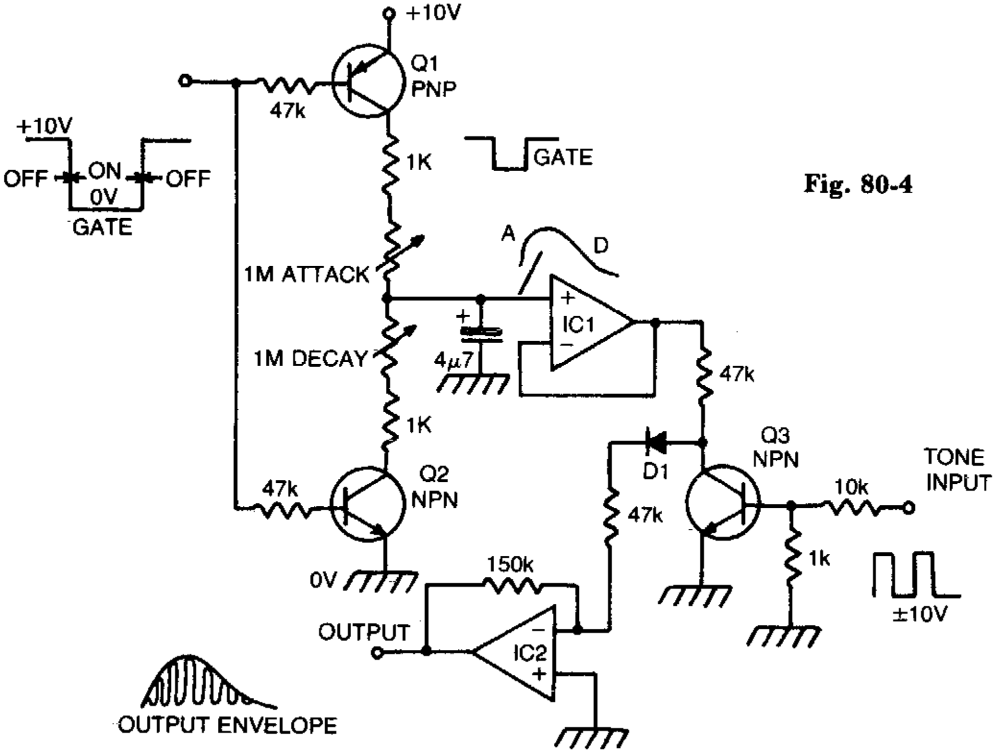

+10V
Q1
PNP
47k
+10V
OFF
ON
OFF
0V
GATE
1K
GATE
1M ATTACK
A
D
+
IC1
−
+
4μ7
1M DECAY
47k
1K
D1
Q3
NPN
TONE
INPUT
10k
Q2
NPN
47k
47k
1k
0V
150k
OUTPUT
−
IC2
+
±10V
OUTPUT ENVELOPE

Fig. 80-4

Circuit Notes

When a gate voltage is applied, Q1 is turned on and capacitor C is charged via the attack pot in series with the 1 K resistor varying this pot, attact time constant. A fast attack gives a percussive sound, a slow attack the affect of "backward" sounds. When the gate voltage returns to its off state, Q2 is turned on and capacitor is discharged via decay pot to ground. The envelope is buffered by IC1 and applied to Q3, which is used as a transistor chopper. A musical tone in the form of a squarewave is connected to the base of Q3. This turns the transistor on or off and thus the envelope is chopped up at regular intervals, the intervals being determined by the pitch of the squarewave. The resultant waveform has the amplitude of the envelope and the harmonic structure of the squarewave. IC2 buffers the signal and D1 ensures that the envelope dies away at the end of a note.

STEREO REVERB SYSTEM

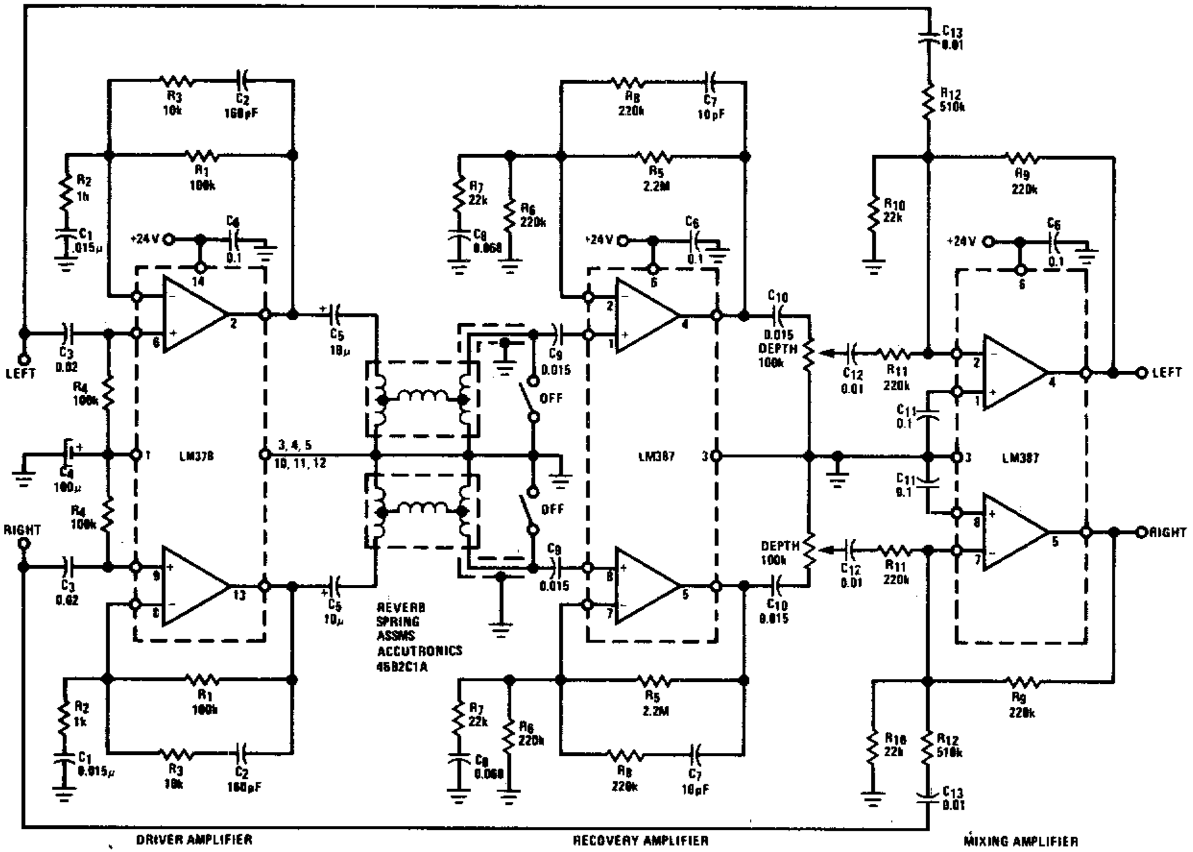

LEFT
RIGHT
+24V
LM378
LM387
C1 .015μ
C1 0.015μ
R2 1k
R3 10k
C2 160pF
R1 100k
C4 0.1
C3 0.02
R4 100k
C4 100μ
3, 4, 5
10, 11, 12
C5 10μ
REVERB SPRING ASSMS ACCUTRONICS 4EB2C1A
OFF
C9 0.015
R7 22k
C8 0.068
R6 220k
R8 220k
C7 10pF
R5 2.2M
C6 0.1
C10 0.015
DEPTH 100k
C12 0.01
R11 220k
C11 0.1
R10 22k
R9 220k
R12 510k
C13 0.01
DRIVER AMPLIFIER
RECOVERY AMPLIFIER
MIXING AMPLIFIER

Fig. 80-5

Circuit Notes

The LM378 dual power amplifier is used as the spring driver. The recovery amplifier is a low noise dual preamplifier. Mixing of the delayed signal with the original is done with another LM387 used in an inverting summing configuration.

FOUR CHANNEL SYNTHESIZER

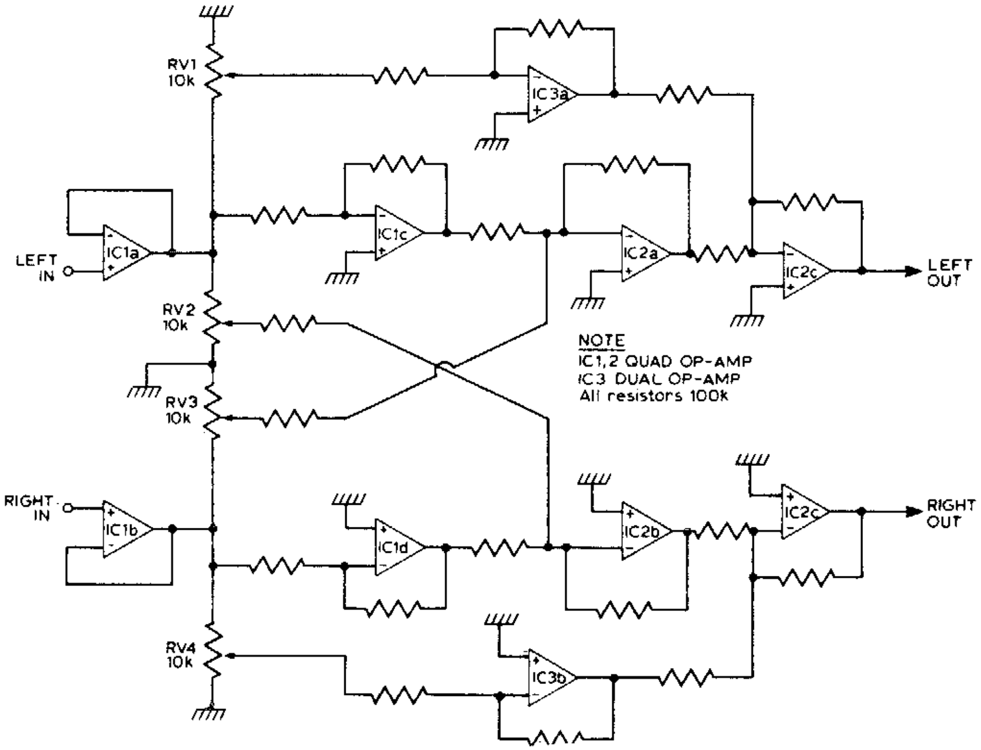
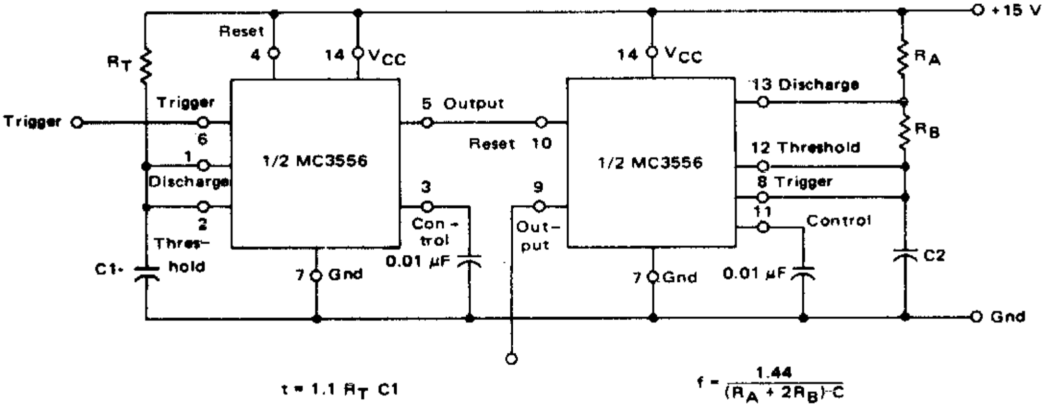

RV1
10k
IC3a
IC1c
IC2a
IC2c
LEFT IN
IC1a
LEFT OUT
RV2
10k
NOTE
IC1,2 QUAD OP-AMP
IC3 DUAL OP-AMP
All resistors 100k
RV3
10k
RIGHT IN
IC1b
IC1d
IC2b
IC2c
RIGHT OUT
RV4
10k
IC3b

Fig. 80-6

Circuit Notes

This circuit will synthesize two rear channels for quadraphonic sound when fed with a stereo signal. The rear output for the left channel, is a combination of the left channel input 180 out of phase, added to a proportion of the right hand channel (also out of phase). The right hand rear output is obtained in a similar way.

TONE BURST GENERATOR



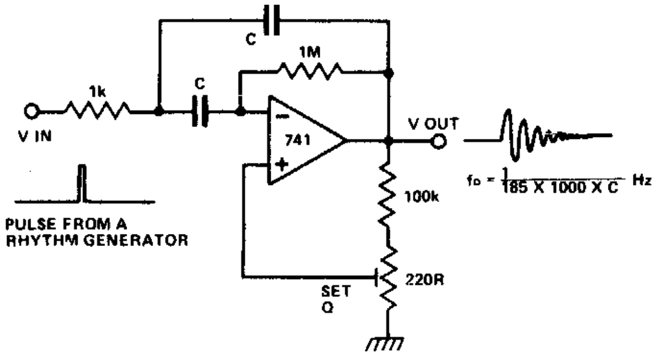
+15 V
Reset
4
14
VCC
R_T
Trigger
Trigger
6
1
Discharge
2
Thres-hold
C1
1/2 MC3556
7
Gnd
5 Output
3
Con-trol
0.01 µF
Reset 10
9
Out-put
1/2 MC3556
14
VCC
13 Discharge
12 Threshold
8 Trigger
11
Control
7
Gnd
0.01 µF
R_A
R_B
C2
Gnd
$t = 1.1\ R_T\ C1$
$f = \frac{1.44}{(R_A + 2R_B) C}$

Fig. 80-7

Circuit Notes

The first timer is used as a monostable and determines the tone duration when triggered by a positive pulse at pin 6. The second timer is enabled by the high output of the monostable. It is connected as an astable and determines the frequency of the tone.

MUSICAL CHIME GENERATOR



C
1M
1k
C
V IN
741
V OUT
PULSE FROM A RHYTHM GENERATOR
100k
220R
SET Q
$f_o = \frac{1}{185 \times 1000 \times C}$ Hz

Fig. 80-8

Circuit Notes

The circuit is that of a multiple feedback bandpass filter. A short click (pulse), makes it ring with a frequency which is its natural resonance frequency. Oscillations die away exponentially and closely resemble many naturally occuring percussive or plucked sounds. The higher the Q the longer the decay time constant. High frequency resonances resemble chimes, lower frequencies sound like claves or bongos. Several circuits, all with different tuning, driven by pulses from a rhythm generator can produce an interesting pattern of sounds.

SOUND EFFECT GENERATOR

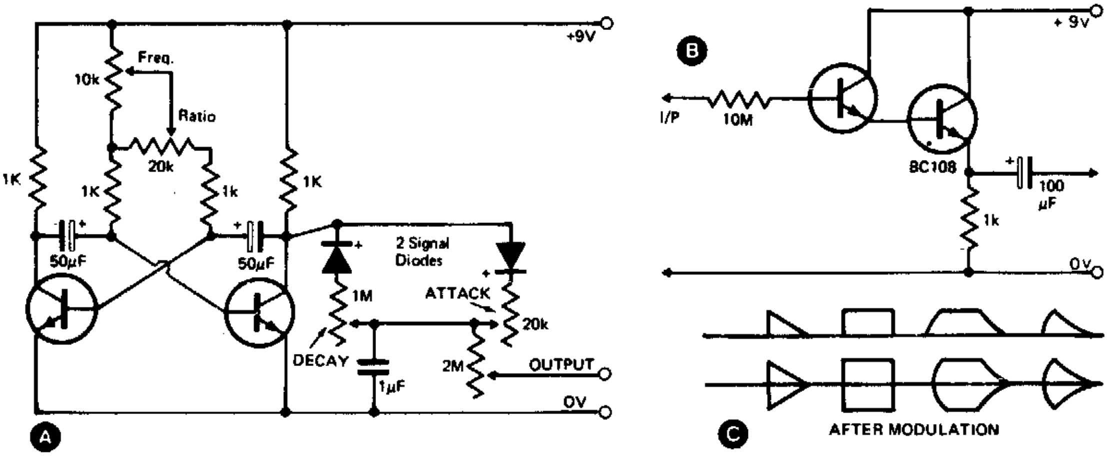

+9V
10k
Freq.
Ratio
20k
1K
1K
1k
1K
50μF
50μF
2 Signal Diodes
1M
ATTACK
20k
DECAY
1μF
2M
OUTPUT
0V
A
B
I/P
10M
BC108
100 μF
1k
+ 9V
0V
C
AFTER MODULATION

Fig. 80-9

Circuit Notes

This waveshape generator is basically a slow running oscillator with variable attack and decay. A variable amplitude (high impedance) output is available via the 2 M potentiometer. B shows an add-on circuit which should be used if a low impedance output is required. Some of the output waveforms that can be produced are shown in C.

PROGRAMMABLE BIRD SOUNDS

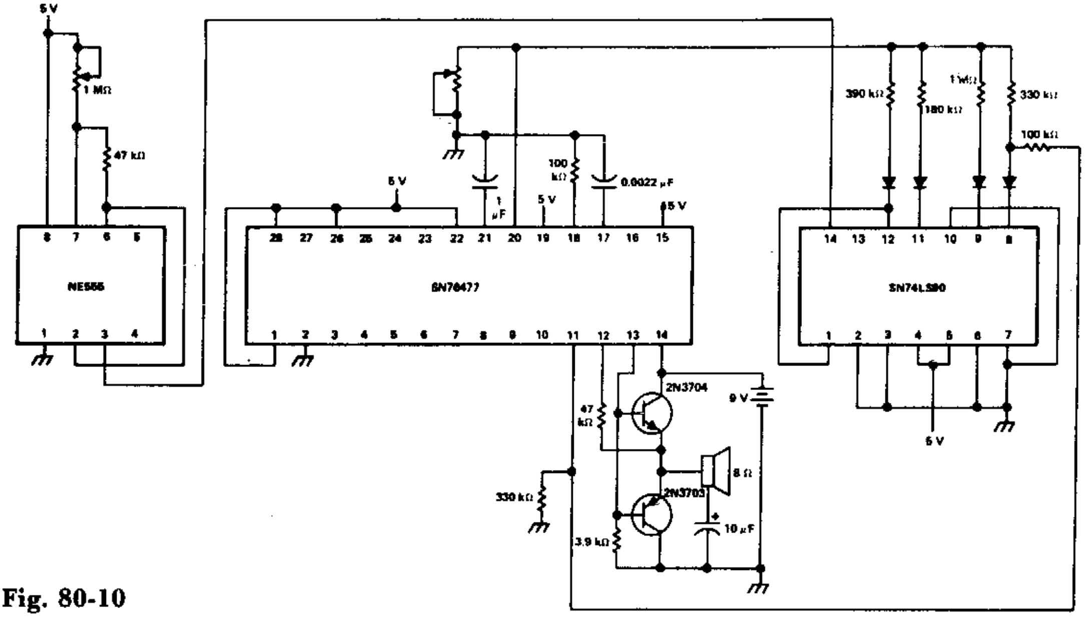

5 V
1 MΩ
47 kΩ
NE555
5 V
1 μF
100 kΩ
0.0022 μF
5 V
5 V
SN76477
390 kΩ
180 kΩ
1 MΩ
330 kΩ
100 kΩ
SN74LS90
5 V
2N3704
9 V
47 kΩ
330 kΩ
8 Ω
2N3703
10 μF
3.9 kΩ

Fig. 80-10

STEREO REVERB ENHANCEMENT SYSTEM

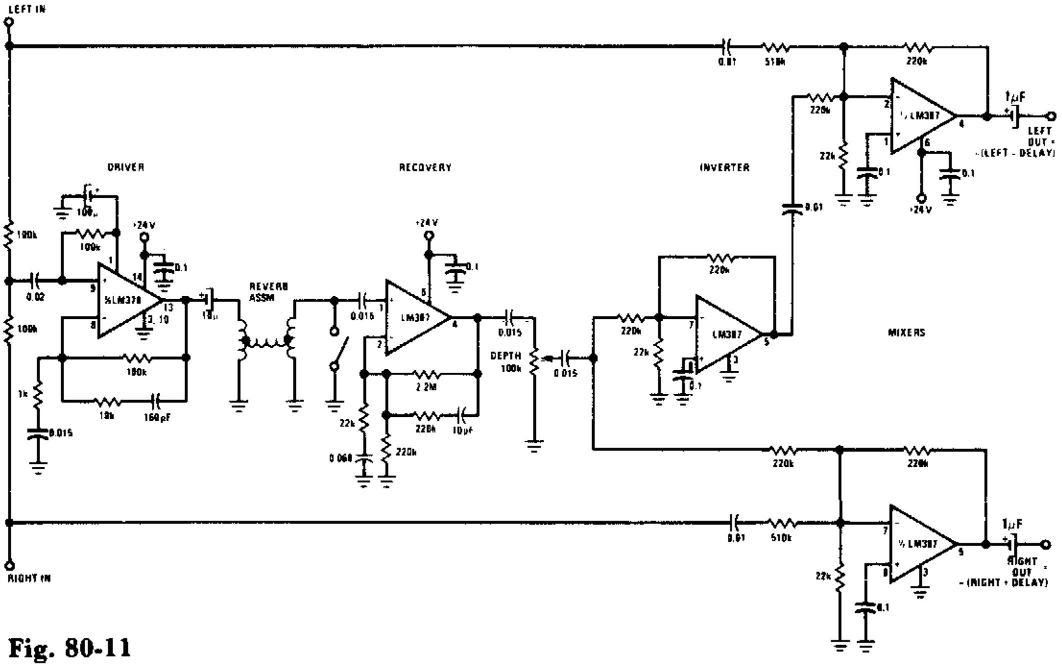

LEFT IN
RIGHT IN
DRIVER
RECOVERY
INVERTER
MIXERS
REVERB ASSM
½LM378
LM387
½ LM387
+24V
100k
100k
0.02
100µ
0.1
10µ
180k
1k
0.015
10k
160pF
0.015
22k
0.068
220k
2.2M
10pF
DEPTH 100k
0.01
510k
220k
22k
1µF
LEFT OUT = (LEFT - DELAY)
RIGHT OUT = (RIGHT + DELAY)

Fig. 80-11

Circuit Notes

The system can be used to synthesize a stereo effect from a monaural source such as AM radio or FM-mono broadcast, or it can be added to an existing stereo (or quad) system where it produces an exciting "opening up" special effect that is truly impressive.

SIREN/SPACE WAR/PHASOR GUN

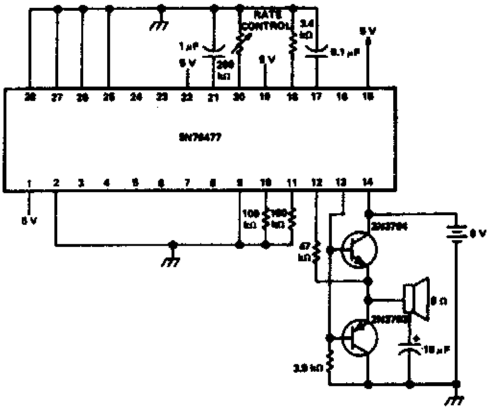

SN76477
28
27
26
25
24
23
22
21
20
19
18
17
16
15
1
2
3
4
5
6
7
8
9
10
11
12
13
14
RATE CONTROL
1 µF
5 V
200 kΩ
3.4 kΩ
0.1 µF
100 kΩ
100 kΩ
47 kΩ
2N3704
2N3702
8 Ω
10 µF
3.9 kΩ
9 V

Fig. 80-12

Circuit Notes

The one shot and decay functions could be added to make an ideal phasor gun sound.

81

Sound (Audio) Operated Circuits

The sources of the following circuits are contained in the Sources section beginning on page 730. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

Voice Activated Switch and Amplifier
Audio Operated Relay
Sound-Modulated Light Source
Audio-Controlled Lamp
Sound Activated Relay
Sound Operated Two-Way Switch

VOICE ACTIVATED SWITCH AND AMPLIFIER

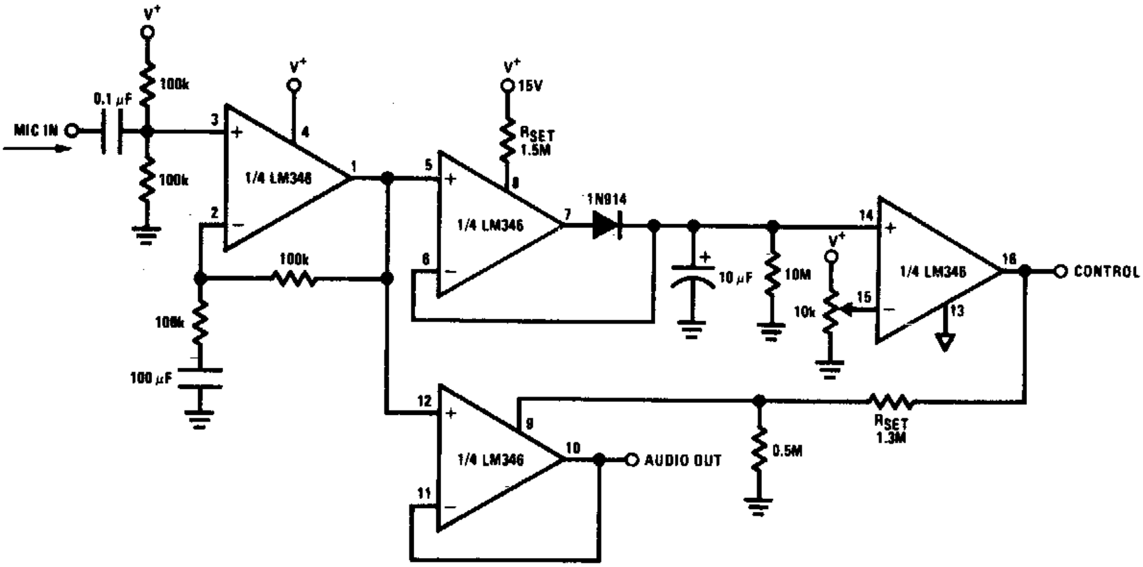

V+
0.1 µF
MIC IN
100k
100k
3
2
1/4 LM346
4
1
V+
V+
15V
RSET
1.5M
5
8
6
1/4 LM346
7
1N914
100k
100k
100 µF
10 µF
10M
V+
10k
14
15
1/4 LM346
16
13
CONTROL
12
9
1/4 LM346
10
11
AUDIO OUT
0.5M
RSET
1.3M

Fig. 81-1

AUDIO OPERATED RELAY

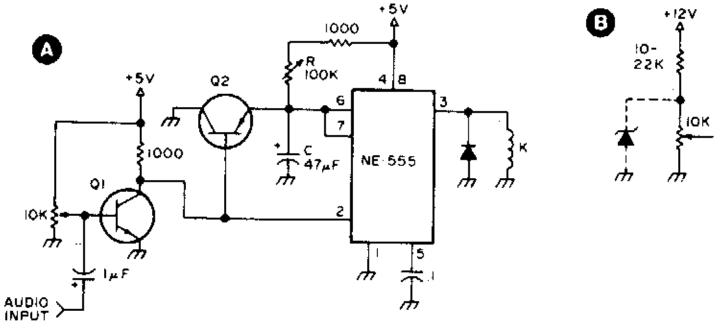
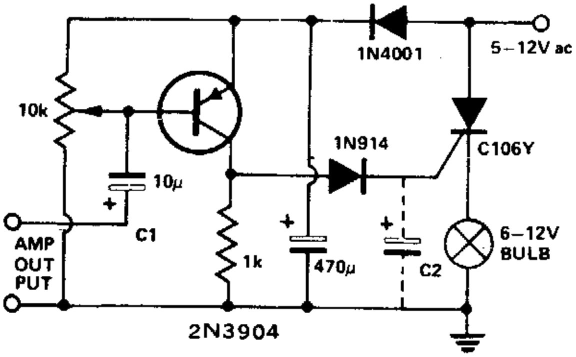

A
+5V
1000
R
100K
4
8
6
7
3
+5V
Q2
C
47µF
NE 555
K
1000
Q1
10K
2
1
5
.1
1µF
AUDIO INPUT
B
+12V
10-
22K
10K

Fig. 81-2

Circuit Notes

Q1 and Q2 are general purpose transistors. The 10 K input pot is adjusted to a point just short of where Q1 turns on as indicated by K pulling in. K is any 5 V reed relay. With the values shown for R (100 K) and C (47 µF), timing values from .05 to slightly over 5 seconds can be achieved. B shows the addition of a 22 K series resistor to the 10 K input pot if a 12 V supply is used. A suitable 12 V reed relay must be used at K.

SOUND-MODULATED LIGHT SOURCE


1N4001
5–12V ac
10k
1N914
C106Y
10µ
+
C1
AMP
OUT
PUT
1k
+
470µ
+
C2
6–12V
BULB
2N3904

Circuit Notes

This circuit modulates a light beam with voice or music from the output of an amplifier. If the 10 K pot is adjusted to slightly less than the V_{be} of the transistor, the circuit forms a peak detector. This drives the gate of the SCR, lighting the bulb whose brightness will vary as the sound level varies. C2 may be removed for a faster response.

Fig. 81-3

AUDIO-CONTROLLED LAMP

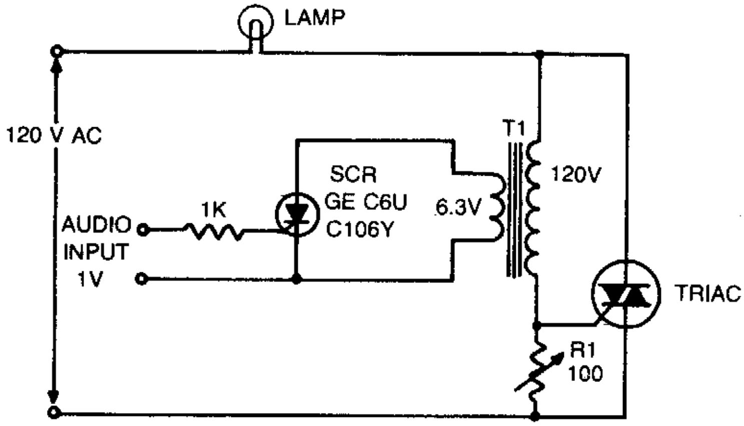

LAMP
120 V AC
T1
SCR
GE C6U
C106Y
6.3V
120V
AUDIO
INPUT
1V
1K
TRIAC
R1
100

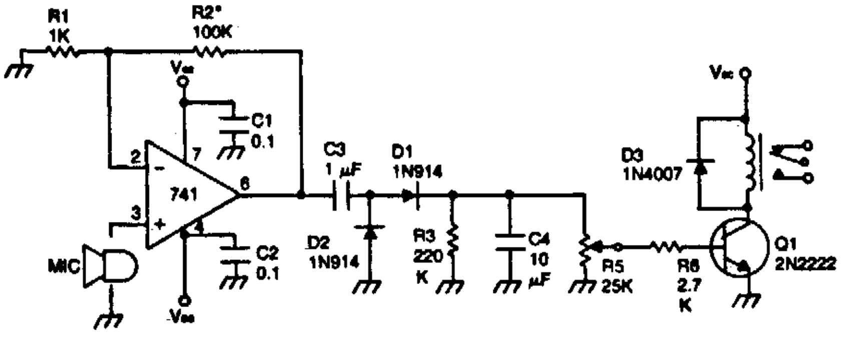
Fig. 81-4

NOTE: T1 IS A 6.3V, 1A. "FILAMENT" TRANSFORMER. ADJUST R1 FOR MAXIMUM RESISTANCE THAT WILL NOT TURN ON LAMP WITH ZERO INPUT.

Circuit Notes

This is an on-off control with isolated, low voltage input. Since the switching action is very rapid, compared with the response time of the lamp and the response of the eye, the effect produced with audio input is similar to a proportional control circuit. If the input signal to the SCR consists of phase-controlled pulses, full wave control of the lamp load is obtained.

SOUND ACTIVATED RELAY


R1
1K
R2*
100K
Vcc
C1
0.1
2
7
741
6
3
4
C2
0.1
MIC
-Vee
C3
1 μF
D1
1N914
D2
1N914
R3
220
K
C4
10
μF
R5
25K
R6
2.7
K
D3
1N4007
Vcc
Q1
2N2222

*SEE TEXT

Fig. 81-5

Circuit Notes

The device remains dormat (in an off condition) until some sound causes it to turn on. The input stage is a 741 operational amplifier connected as a noninverting follower audio amplifier. Gain is approximately 100. To increase gain raise the value of R2. The amplified signal is rectified and filtered to a dc level by R4. Then R5 is set to the audio level desired to activate the relay.

SOUND OPERATED TWO-WAY SWITCH

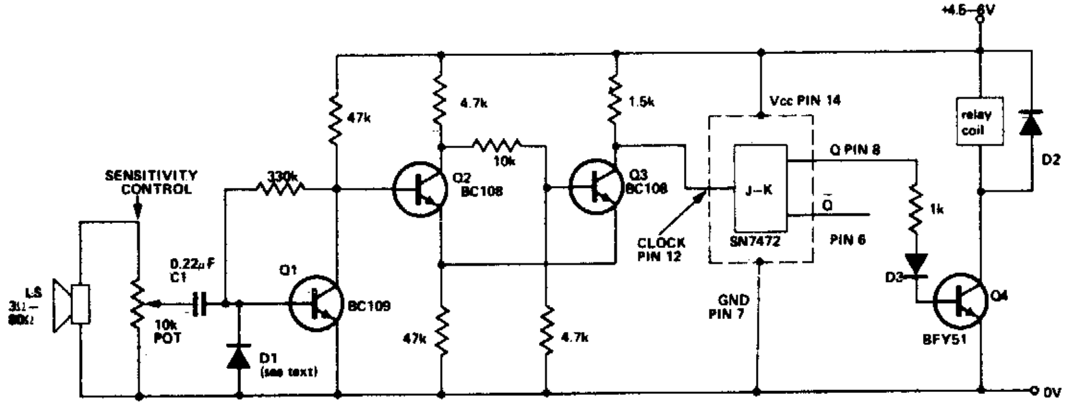

+4.5–6V
47k
4.7k
1.5k
Vcc PIN 14
relay coil
D2
10k
Q PIN 8
SENSITIVITY CONTROL
330k
Q2
BC108
Q3
BC108
J–K
Q̄
1k
CLOCK PIN 12
SN7472
PIN 6
0.22μF
C1
Q1
D3
Q4
LS
3Ω–80Ω
10k POT
BC109
D1
(see text)
47k
4.7k
GND PIN 7
BFY51
0V

Fig. 81-6

Circuit Notes

This circuit operates a relay each time a sound of sufficient intensity is made, thus one clap of the hands will switch it one way, a second clap will revert the circuit to the original condition. Q2 and Q3 form a Schmitt trigger. The JK flip-flop is used as a bistable whose output changes state every time a pulse is applied to the clock input (pin 12). Q4 allows the output to drive a relay.

82

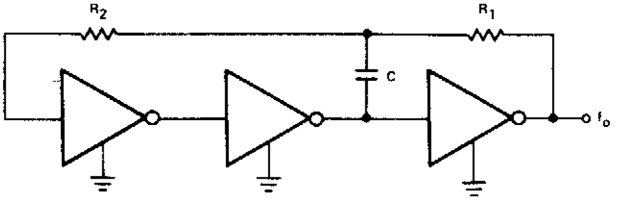
Square Wave Oscillators

The sources of the following circuits are contained in the Sources section beginning on page 730. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

R/C Oscillator
1 kHz Square Wave Oscillator
TTL Oscillator
Square Wave Oscillator
Adjustable TTL Clock
Square Wave Oscillator
Oscillator/Clock Generator
CMOS Oscillator
Free-Running Square-Wave Oscillator
Precision Squares
Square Wave Oscillator
0.5 Hz Square-Wave Oscillator
Simple Triangle/Square Wave Oscillator
Squarewave Oscillator

R/C OSCILLATOR

$$f_O \approx \frac{1}{2\,C[0.41\,R_P + 0.70\,R_1]}\quad, R_P = \frac{R_1\,R_2}{R_1 + R_2}$$



Gates are 74C04

a. If $R_1 = R_2 = R_1$, $f \cong 0.55/RC$

b. If $R_2 >> R_1$, $f \cong 0.45/R_1C$

c. If $R_2 << R_1$, $f \cong 0.72/R_1C$

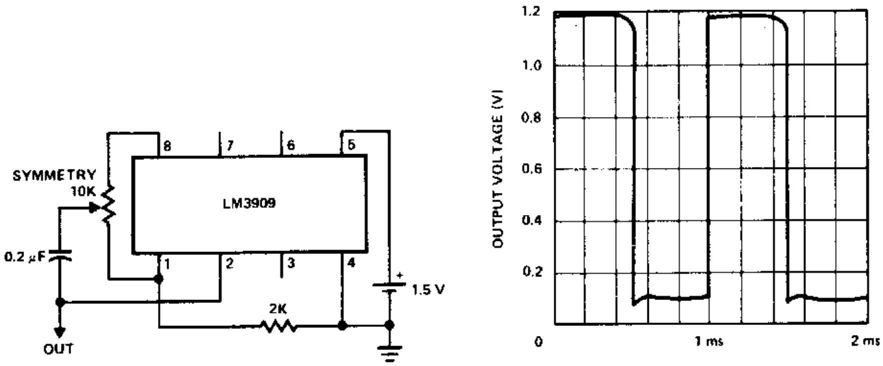
a. $f = 120$ kHz, $C = 420$ pF
$R_1 = R_2 \approx 10.9\ \text{k}\,\Omega$

b. $f = 120$ kHz, $C = 420$ pF, $R_2 = 50\ \text{k}\,\Omega$
$R_1 = 8.93\ \text{k}\,\Omega$

c. $f = 120$ kHz, $C = 220$ pF, $R_2 = 5\ \text{k}\,\Omega$
$R_1 = 27.3\ \text{k}\,\Omega$

Fig. 82-1

1 kHz SQUARE WAVE OSCILLATOR


8
7
6
5
SYMMETRY
10K
LM3909
0.2 µF
1
2
3
4
2K
1.5 V
OUT
1.2
1.0
0.8
0.6
0.4
0.2
0
OUTPUT VOLTAGE (V)
1 ms
2 ms

Note: Output Voltage Through a 10K Load to Ground

Fig. 82-2

TTL OSCILLATOR

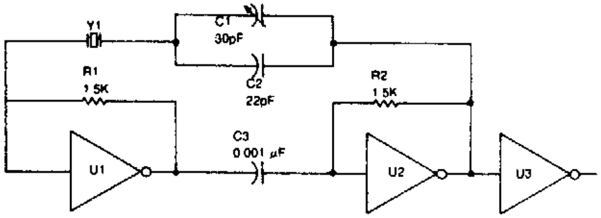

Y1
C1
30pF
C2
22pF
R1
1 5K
R2
1 5K
C3
0 001 μF
U1
U2
U3

Fig. 82-3

Circuit Notes

TTL inverter stages, U1 and U2, are cross-connected with a crystal Y1. A resistor in each stage biases the normally digital gates into a region where they operate as amplifiers. Inverter stage U3 is used as a buffer.

SQUARE WAVE OSCILLATOR

Oscillator Frequency for Various Capacitor Values

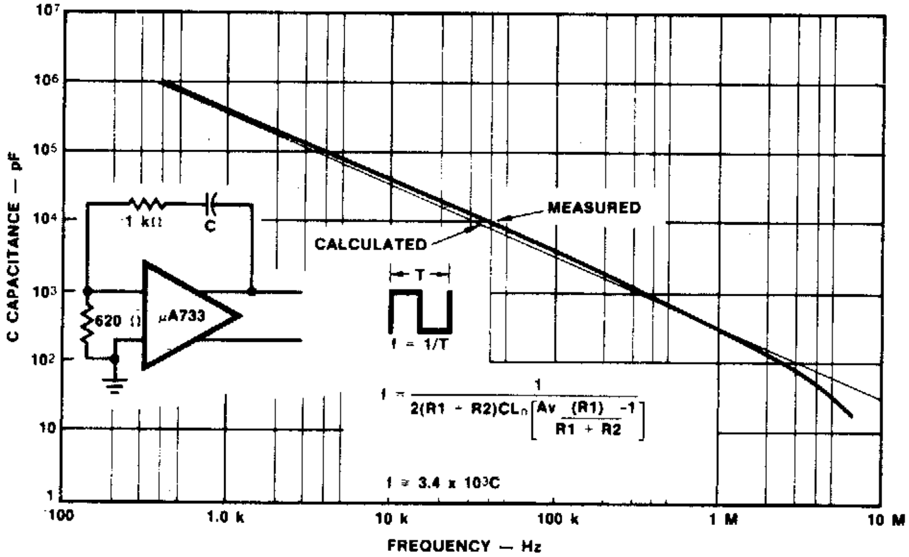


Fig. 82-4

ADJUSTABLE TTL CLOCK (MAINTAINS 50% DUTY CYCLE)

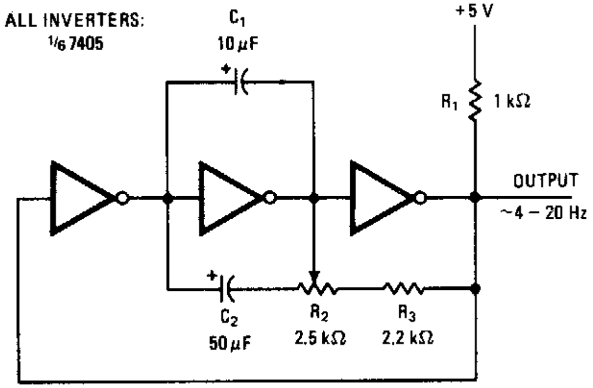

ALL INVERTERS:
1/6 7405
C1
10 μF
+5 V
R1
1 kΩ
OUTPUT
~4 – 20 Hz
C2
50 μF
R2
2.5 kΩ
R3
2.2 kΩ

Fig. 82-5

Circuit Notes

Symmetry of the square-wave output is maintained by connecting the right side of R2 through resistor R3 to the output of the third amplifier stage. This changes the charging current to the capacitors in proportion to the setting of frequency-adjusting potentiometer R2. Thus, a duty cycle of 50% is constant over the entire range of oscillation. The lower frequency limit is set by capacitor C2. With the components shown, the frequency of oscillation can be varied by R2 from about 4 to 20 hertz. Other frequency ranges can be obtained by changing the values of C1 and R3, which control the upper limit of oscillation, or C2, which limits the low-frequency end.

SQUARE WAVE OSCILLATOR

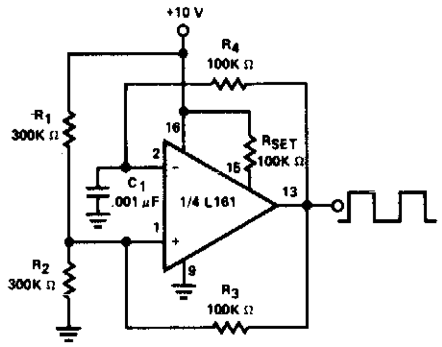
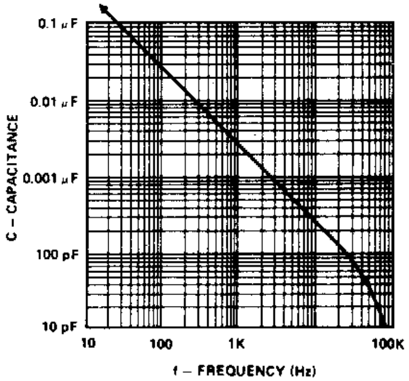

+10 V
R4
100K Ω
R1
300K Ω
16
RSET
100K Ω
2
15
C1
.001 μF
1/4 L161
13
1
9
R2
300K Ω
R3
100K Ω

Fig. 82-6



Frequency vs the Value of C_1 for the Squarewave Oscillator

Circuit Notes

This generator is operable to over 100 kHz. The low frequency limit is determined by C1. Frequency is constant for supply voltages down to +5 V.

OSCILLATOR/CLOCK GENERATOR

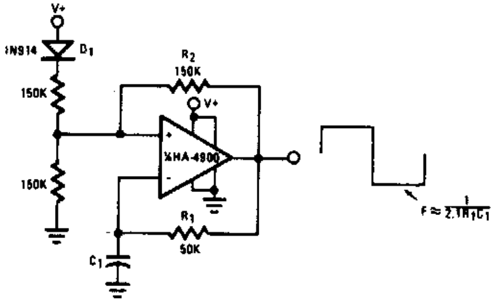
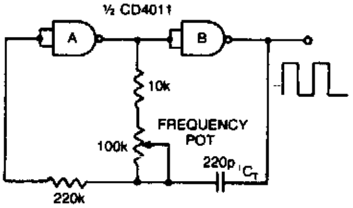

V+
1N914
D1
150K
R2
150K
V+
¼HA-4900
150K
R1
50K
C1
$F \approx \frac{1}{2.1R_1C_1}$

Fig. 82-7

Circuit Notes

This self-starting fixed frequency oscillator circuit gives excellent frequency stability. R1 and C1 comprise the frequency determining network while R2 provides the regenerative feedback. Diode D1 enhances the stability by compensating for the difference between V_{OH} and V_{Supply}. In applications where a precision clock generator up to 100 kHz is required, such as in automatic test equipment, C1 may be replaced by a crystal.

CMOS OSCILLATOR


½ CD4011
A
B
10k
FREQUENCY
POT
100k
220p
C_T
220k

Circuit Notes

Varying the 100 K pot changes the discharge rate of C_T and hence the frequency. A square wave output is generated. The maximum frequency using CMOS is limited to 2 MHz.

Fig. 82-8

FREE-RUNNING SQUARE-WAVE OSCILLATOR

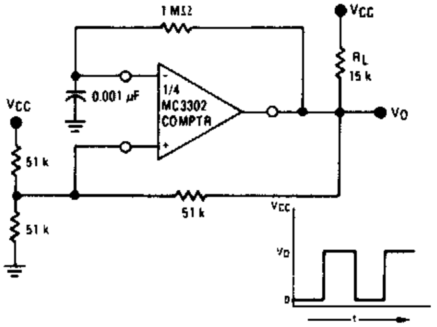
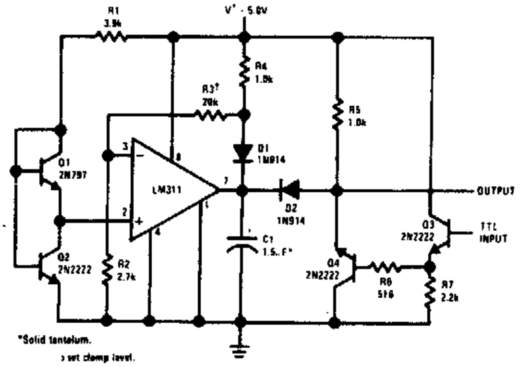

1 MΩ
Vcc
RL
15 k
0.001 μF
1/4
MC3302
COMPTR
Vcc
Vo
51 k
51 k
51 k
Vcc
Vo
0
t

Fig. 82-9

PRECISION SQUARER


R1
3.9k
V⁺ = 5.0V
R4
1.0k
R3†
20k
R5
1.0k
Q1
2N797
D1
1N914
3
8
7
LM311
2
1
4
D2
1N914
OUTPUT
Q3
2N2222
TTL
INPUT
C1
1.5μF*
Q2
2N2222
R2
2.7k
Q4
2N2222
R6
510
R7
2.2k

*Solid tantalum.

† set clamp level.

Fig. 82-10

SQUARE WAVE OSCILLATOR

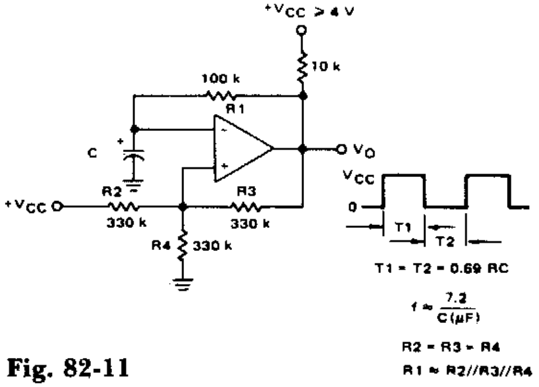

+Vcc > 4 V
10 k
100 k
R1
C
+
R2
+Vcc
330 k
R3
330 k
R4
330 k
Vo
Vcc
0
T1
T2
T1 = T2 = 0.69 RC
f ≈ 7.2/C(μF)
R2 = R3 = R4
R1 ≈ R2//R3//R4

Fig. 82-11

0.5 Hz SQUARE-WAVE OSCILLATOR

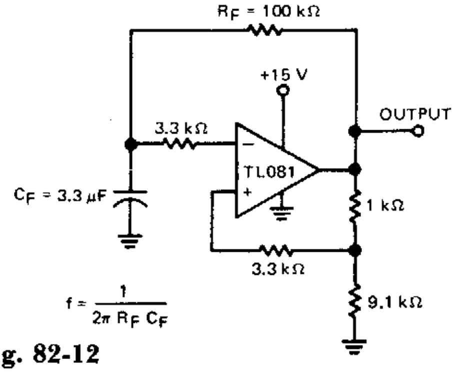

RF = 100 kΩ
+15 V
OUTPUT
3.3 kΩ
TL081
CF = 3.3 μF
1 kΩ
3.3 kΩ
9.1 kΩ
f = 1/(2π RF CF)

Fig. 82-12

SIMPLE TRIANGLE/SQUARE WAVE OSCILLATOR

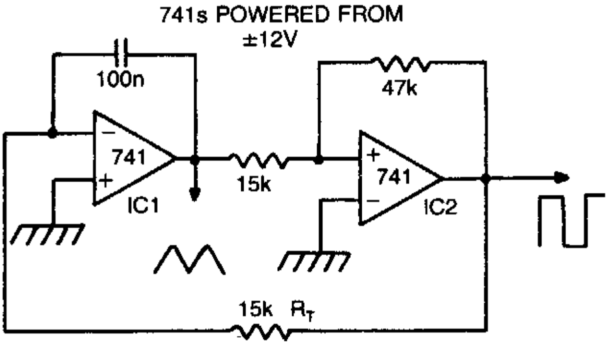

741s POWERED FROM ±12V
100n
47k
741
IC1
15k
741
IC2
15k RT

Fig. 82-13

Circuit Notes

By making RT variable it is possible to alter the operating frequency over a 100 to 1 range. Versatile triangle/square-wave oscillator has a possible frequency range of 0.1 Hz to 100 kHz.

SQUAREWAVE OSCILLATOR

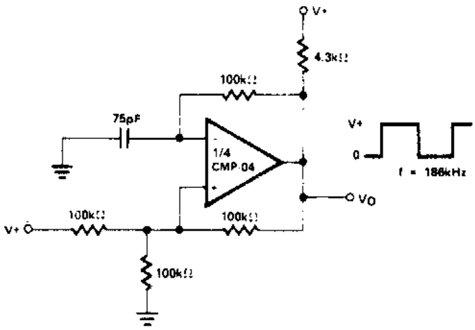

V+
4.3kΩ
100kΩ
75pF
1/4 CMP-04
V+
0
f = 186kHz
Vo
V+
100kΩ
100kΩ
100kΩ

Fig. 82-14

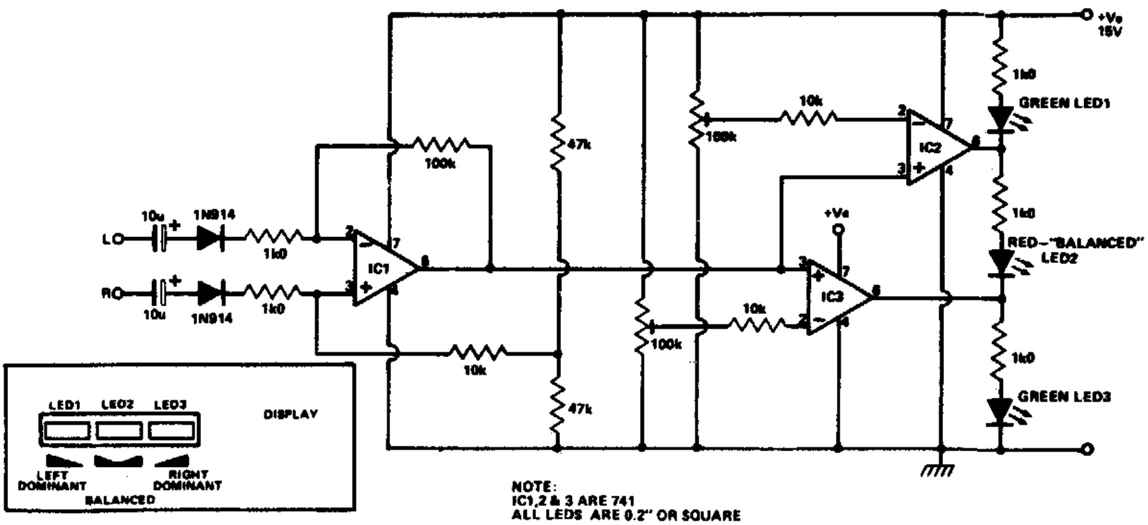
83

Stereo Balance Circuits

The sources of the following circuits are contained in the Sources section beginning on page 730. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

Stereo Balance Meter
Stereo Balancer
Stereo Balance Meter

STEREO BALANCE METER


+Vs 15V
1k0
GREEN LED1
10k
100k
IC2
47k
100k
10u
1N914
L
R
1k0
IC1
+Vs
RED-"BALANCED" LED2
IC3
10k
100k
10k
47k
GREEN LED3
LED1 LED2 LED3
DISPLAY
LEFT DOMINANT
RIGHT DOMINANT
BALANCED

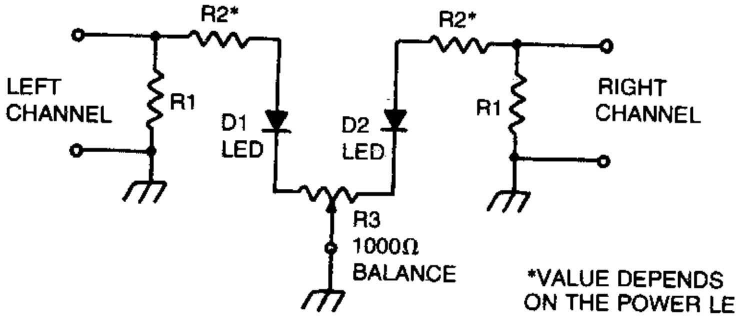
NOTE:
IC1,2 & 3 ARE 741
ALL LEDS ARE 0.2" OR SQUARE

Fig. 83-1

Circuit Notes

Outputs from each channel are fed to the two inputs of IC1 connected as a differential amplifier. IC2 and 3 are driven by the output of IC1. Output of IC1 is connected to the noninverting inputs of IC2 and 3. If the output of IC1 approaches the supply rail, the outputs of ICs 2 and 3 will also go high, illuminating LED3. This would happen if the right channel were dominating. If the left channel was dominant, the outputs of ICs 2 and 3 would be low, illuminating LED1. If the two channels are equal in amplitude, the outputs of ICs 2 and 3 would be high and low respectively, lighting up LED2.

STEREO BALANCER



*VALUE DEPENDS
ON THE POWER LEVEL

Fig. 83-2

Circuit Notes

This circuit will allow you to set the gain of two stereo channels to the same level. The signal across the two channel-load resistors is sampled by resistors R2. (Values of these resistors will depend upon the power level.) For most 20 milliampere LED, use approximately 2.5 K per watt. (For a 10-watt system use a 25,000 ohm resistor.) To set up, short the two inputs and connect them to one channel of a power amplifier. Apply a signal and adjust R3 until both LEDs glow at the same brightness level. The balancer is ready for use. Connect the inputs of the stereo balancer across the output of the power amplifier, and then turn up either the independent volume controls, or the balance control until both LEDs glow at the same level. To use this circuit in-line with loudspeakers, disconnect both R1s, and use the speakers as the load.

STEREO BALANCE METER

PARTS LIST FOR
STEREO BALANCE METER

D1, D2—Silicon rectifier rated 100 PIV at any low current
M1—Zero-center DC mA meter (see text)
R1, R2—1000-ohm, ½-watt resistor, 5% or 1%

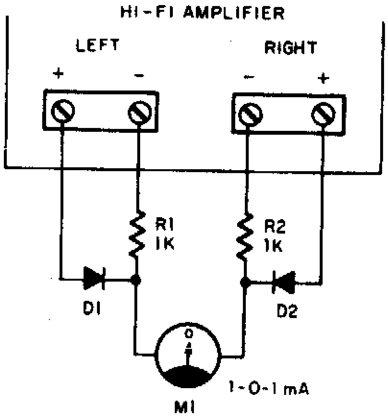


Fig. 83-3

Circuit Notes

Play any stereo disc or tape and then set the amplifier to mono. Adjust left and right channel balance until meter M1 indicates zero; then the left and right output level are identical.

84

Switches

The sources of the following circuits are contained in the Sources section beginning on page 730. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

- DTL-TTL Controlled Buffered Analog Switch
- High Toggle Rate High Frequency Analog Switch
- Differential Analog Switch
- High Frequency Switch
- Two-Channel Switch
- 10 A, 25 VDC Solid State Relays

DTL-TTL CONTROLLED BUFFERED ANALOG SWITCH

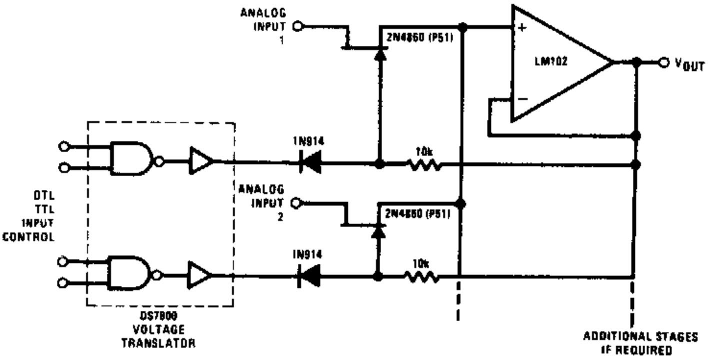

ANALOG INPUT 1
2N4860 (P51)
LM102
+
−
V_{OUT}
1N914
10k
DTL TTL INPUT CONTROL
ANALOG INPUT 2
2N4860 (P51)
1N914
10k
DS7800 VOLTAGE TRANSLATOR
ADDITIONAL STAGES IF REQUIRED

Fig. 84-1

Circuit Notes

This analog switch uses the 2N4860 JFET for its 25 ohm r_{on} and low leakage. The LM102 serves as a voltage buffer. This circuit can be adapted to a dual trace oscilloscope chopper. The DS7800 monolithic IC provides adequate switch drive controlled by DTL/TTL logic levels.

HIGH TOGGLE RATE HIGH FREQUENCY ANALOG SWITCH

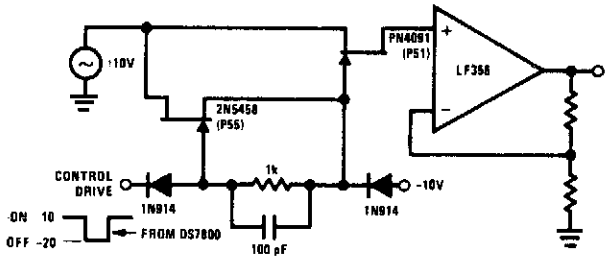
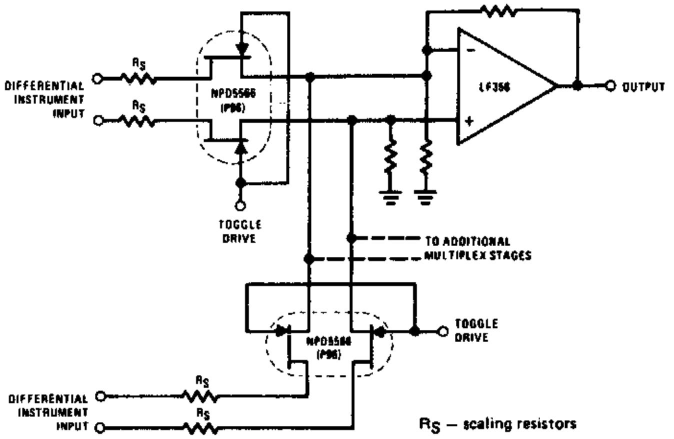

±10V
PN4091 (P51)
+
−
LF356
2N5458 (P55)
CONTROL DRIVE
1N914
1k
100 pF
1N914
−10V
ON 10
OFF −20
FROM DS7800

Fig. 84-2

Circuit Notes

Commutator circuit provides low impedance gate drive to the PN4091 analog switch for both on and off drive conditions. This circuit also approaches the ideal gate drive conditions for high frequency signal handling by providing a low ac impedance for off drive and high ac impedance for on drive to the PN4091

DIFFERENTIAL ANALOG SWITCH


DIFFERENTIAL INSTRUMENT INPUT
R_S
R_S
NPD5566 (P96)
TOGGLE DRIVE
LF356
OUTPUT
TO ADDITIONAL MULTIPLEX STAGES
TOGGLE DRIVE
NPD5566 (P96)
R_S
R_S
DIFFERENTIAL INSTRUMENT INPUT

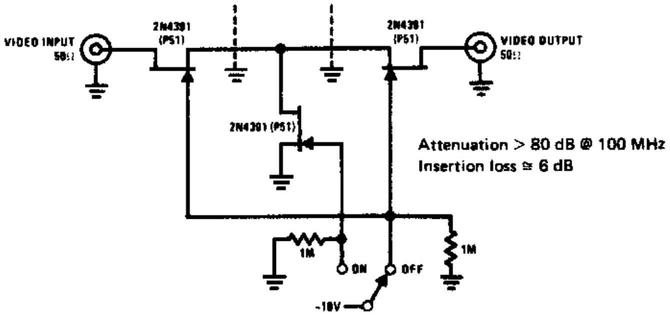
R_S – scaling resistors

Fig. 84-3

Circuit Notes

The NPD5566 monolithic dual is used in a differential multiplex application where $R_{ds(ON)}$ should be closely matched. Since $R_{ds(ON)}$ for the monolithic dual tracks at better than ±1% over wide temperature ranges (−25° C to +125° C), this makes it an unusual but ideal choice for an accurate multiplexer. This close tracking greatly reduces errors due to common-mode signals.

HIGH FREQUENCY SWITCH


VIDEO INPUT 50Ω
2N4391 (P51)
2N4391 (P51)
VIDEO OUTPUT 50Ω
2N4391 (P51)
1M
ON
OFF
1M
-10V

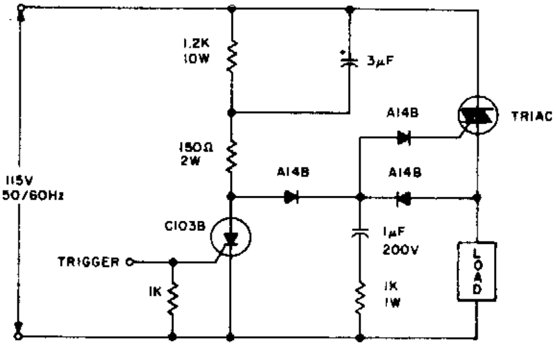
Attenuation > 80 dB @ 100 MHz
Insertion loss ≅ 6 dB

Fig. 84-4

Circuit Notes

The 2N4391 provides a low ON resistance of 30 ohm and a high OFF impedance (< 0.2 pF) when off. With proper layout and an ideal switch, the performance stated above can be readily achieved.

TRIAC ZERO VOLTAGE SWITCHING

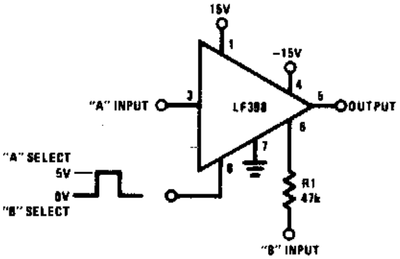

1.2K
10W
3μF
A14B
TRIAC
150Ω
2W
A14B
A14B
115V
50/60Hz
C103B
A14B
1μF
200V
TRIGGER
1K
1K
1W
LOAD

Circuit Notes

The triac will be gated on at the start of the positive half cycle by current flow through the 3 μF capacitor as long as the C103 SCR is off. The load voltage then charges up the 1 μF capacitor so that the triac will again be energized during the subsequent negative half cycle of line voltage. A selected gate triac is required because of the III+ triggering mode.

Fig. 84-5

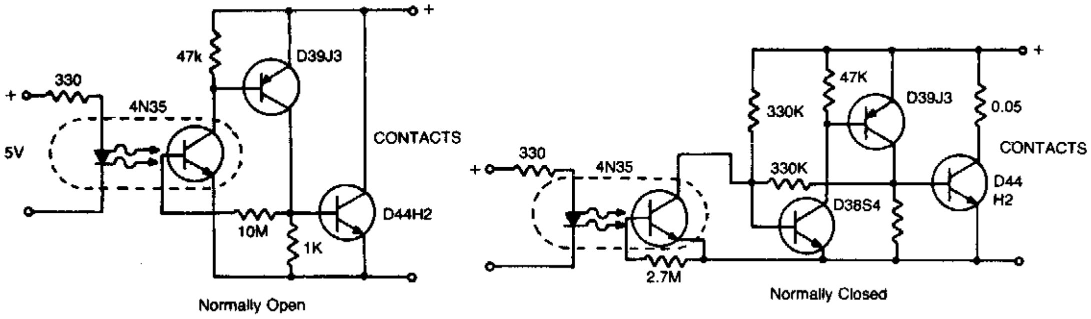
TWO-CHANNEL SWITCH


15V
1
−15V
4
"A" INPUT
3
LF398
5
OUTPUT
6
7
8
"A" SELECT
5V
0V
"B" SELECT
R1
47k
"B" INPUT

	A	B
Gain	1 ±0.02%	1 ±0.2%
Z_{IN}	$10^{10}\Omega$	47 kΩ
BW	≃1 MHz	≈400 kHz
Crosstalk @ 1 kHz	−90 dB	−90 dB
Offset	≤ 6 mV	≤ 75 mV

Fig. 84-6

10 A, 25 Vdc SOLID STATE RELAYS


+
47k
D39J3
330
+
4N35
5V
CONTACTS
10M
1K
D44H2
+
47K
D39J3
330K
0.05
CONTACTS
330
+
4N35
330K
D38S4
D44
H2
2.7M

Normally Open

Normally Closed

Fig. 84-7

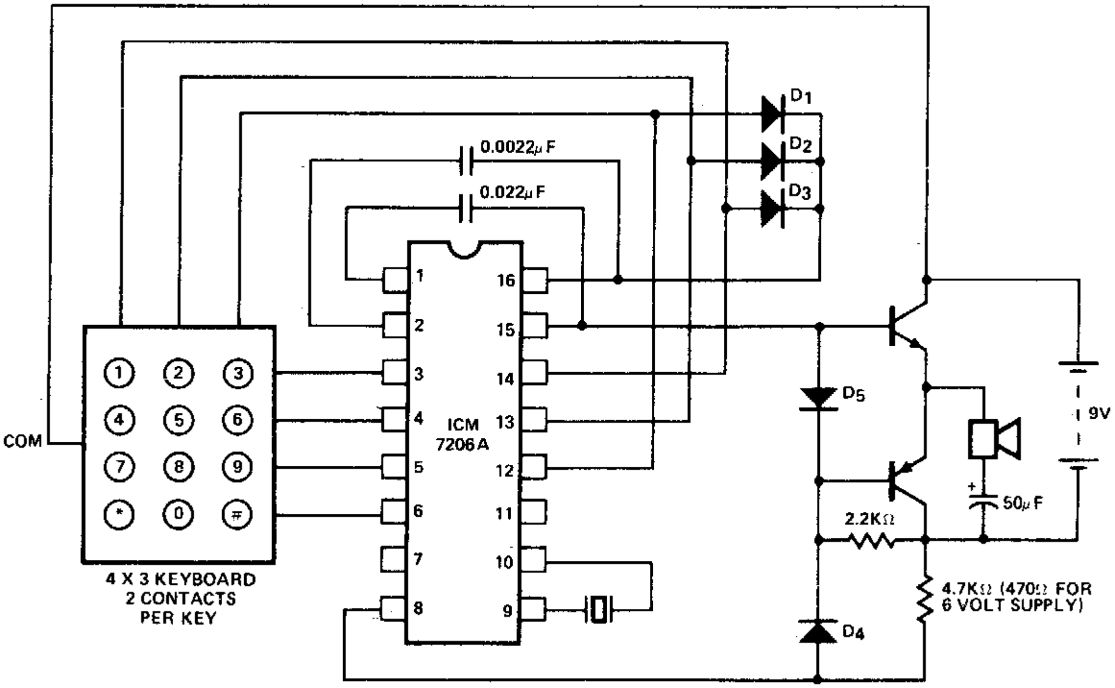
85

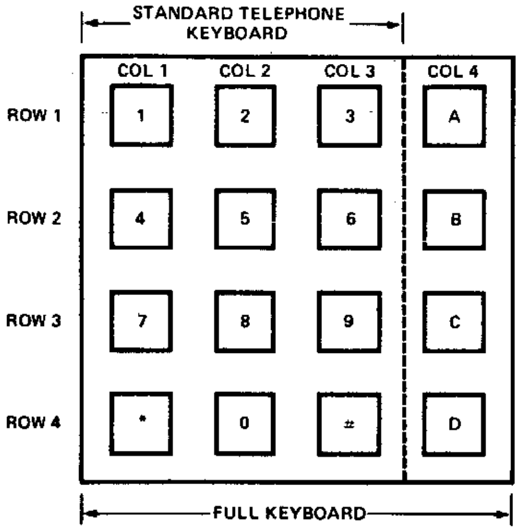
Telephone Related Circuits

The sources of the following circuits are contained in the Sources section beginning on page 730. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

Portable Tone Generator
Telephone Status Monitor Using an Optoisolator
Telephone Tone Ringer
F.C.C. Approved Telephone Tone Ringer
Telephone or Extension Tone Ringer
Telephone Line Monitor
Tone Dial Generator
Tone Dial Encoder
Tone Dial Sequence Decoder
Remote Ring Extender Switch
Tone Dial Decoder
Telephone Relay
Telephone-Controlled Tape Starter (TCTS)
Telephone Line Powered Repertory Dialer
Telephone Off-Hook Indicator
Telephone Handset Tone Dial Encoder
Low Line Loading Ring Detector
Phone Auto Answer and Ring Indicator
Autopatch Telephone Phone Line Interface
Telephone Ringer Uses Piezoelectric Device
Electronic Phone Bell

PORTABLE TONE GENERATOR


D1
D2
D3
0.0022μF
0.022μF
1
2
3
4
5
6
7
8
16
15
14
13
12
11
10
9
ICM
7206A
COM
1 2 3
4 5 6
7 8 9
* 0 #
4 X 3 KEYBOARD
2 CONTACTS
PER KEY
D5
9V
50μF
2.2KΩ
4.7KΩ (470Ω FOR
6 VOLT SUPPLY)
D4


STANDARD TELEPHONE
KEYBOARD
COL 1
COL 2
COL 3
COL 4
ROW 1
ROW 2
ROW 3
ROW 4
1
2
3
A
4
5
6
B
7
8
9
C
*
0
#
D
FULL KEYBOARD

KEY	LOW BAND FREQ. Hz	HI BAND FREQ. Hz
1	697	1209
2	697	1336
3	697	1477
4	770	1209
5	770	1336
6	770	1477
7	852	1209
8	852	1336
9	852	1477
*	941	1209
0	941	1336
#	941	1477
A	697	1633
B	770	1633
C	852	1633
D	941	1633

Fig. 85-1

TELEPHONE STATUS MONITOR USING AN OPTOISOLATOR

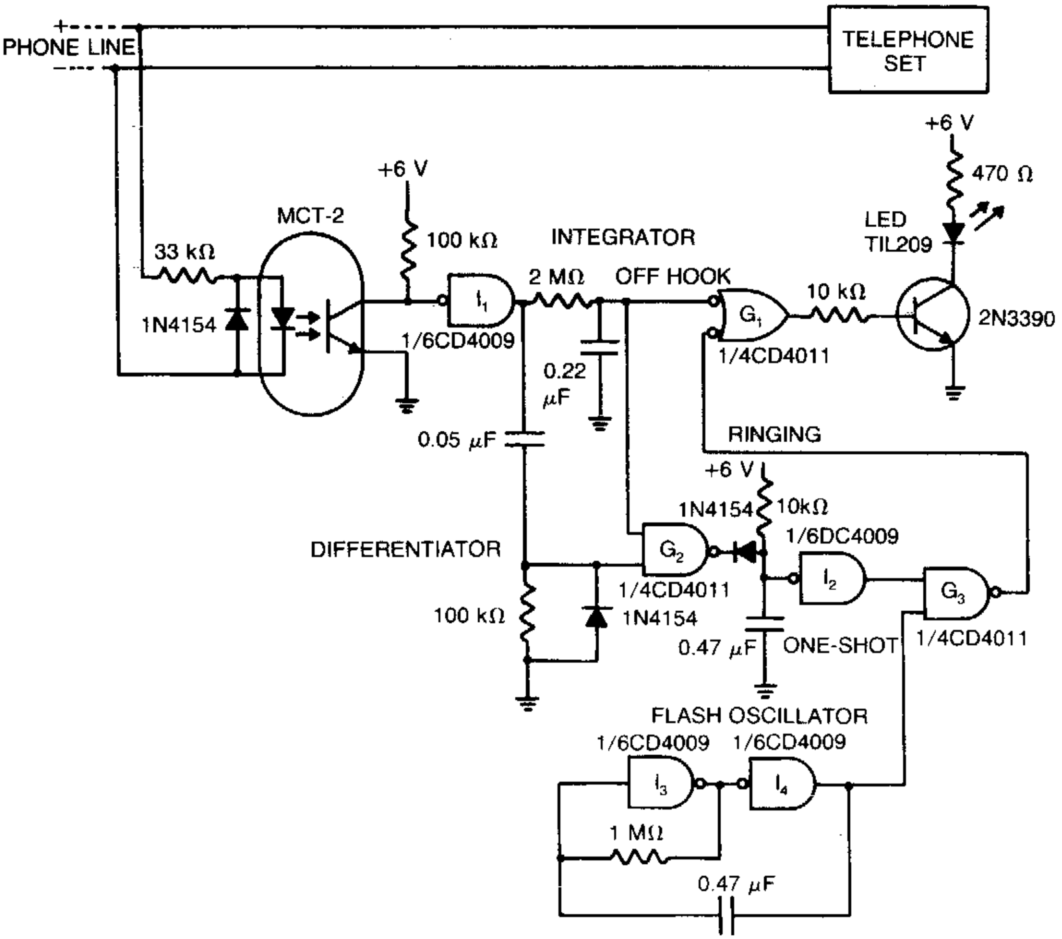
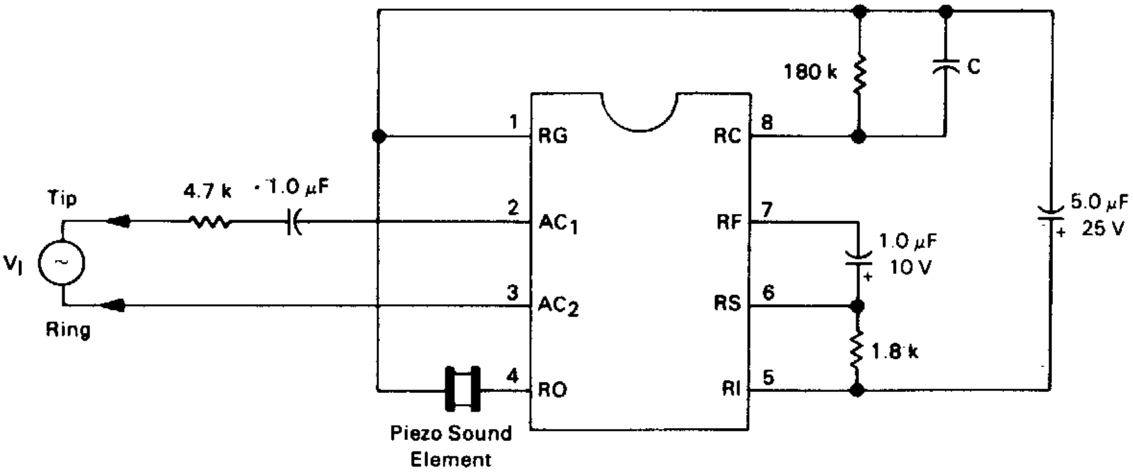

+
PHONE LINE
−
TELEPHONE SET
+6 V
470 Ω
LED TIL209
2N3390
10 kΩ
G1
1/4CD4011
+6 V
100 kΩ
MCT-2
33 kΩ
1N4154
I1
1/6CD4009
INTEGRATOR
2 MΩ
OFF HOOK
0.22 μF
0.05 μF
RINGING
+6 V
1N4154
10kΩ
1/6DC4009
DIFFERENTIATOR
G2
1/4CD4011
1N4154
100 kΩ
I2
G3
0.47 μF
ONE-SHOT
1/4CD4011
FLASH OSCILLATOR
1/6CD4009
1/6CD4009
I3
I4
1 MΩ
0.47 μF

Fig. 85-2

Circuit Notes

The LED indicates the status of a remote telephone. The light is off if the phone is hung up. It shines steadily if the phone is off hook, and it flashes on and off while phone rings and for 5 seconds after ringing stops. The flashing oscillator operates continuously but can drive the LED only when a ringing signal discharges the one shot capacitor to enable NAND gate G3. Thus, one oscillator handles several phone lines.

TELEPHONE TONE RINGER


180 k
C
1
RG
RC
8
Tip
4.7 k
1.0 μF
2
AC$_1$
RF
7
5.0 μF
25 V
V_I
1.0 μF
10 V
3
AC$_2$
RS
6
Ring
1.8 k
4
RO
RI
5
Piezo Sound
Element

MC34012-1: C = 1000 pF
MC34012-2: C = 500 pF
MC34012-3: C = 2000 pF

Fig. 85-3

Circuit Notes

This is a complete telephone bell replacement circuit with minimum external components with on-chip diode bridge and transient protection and direct drive for piezoelectric transducers.

F.C.C. APPROVED TELEPHONE TONE RINGER

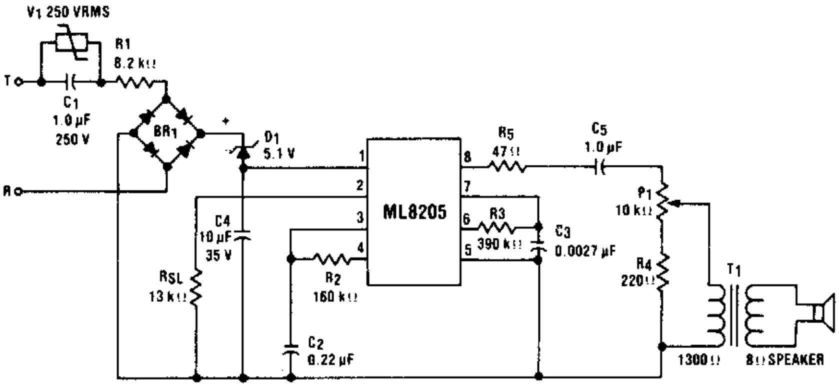

V$_1$ 250 VRMS
R1
8.2 kΩ
T
C$_1$
1.0 μF
250 V
BR$_1$
+
D1
5.1 V
R
ML8205
1
2
3
4
8
7
6
5
R5
47 Ω
C5
1.0 μF
R3
390 kΩ
C3
0.0027 μF
P$_1$
10 kΩ
R4
220 Ω
T1
C4
10 μF
35 V
R$_{SL}$
13 kΩ
R$_2$
160 kΩ
C$_2$
0.22 μF
1300 Ω
8 Ω SPEAKER

Fig. 85-4

TELEPHONE OR EXTENSION TONE RINGER

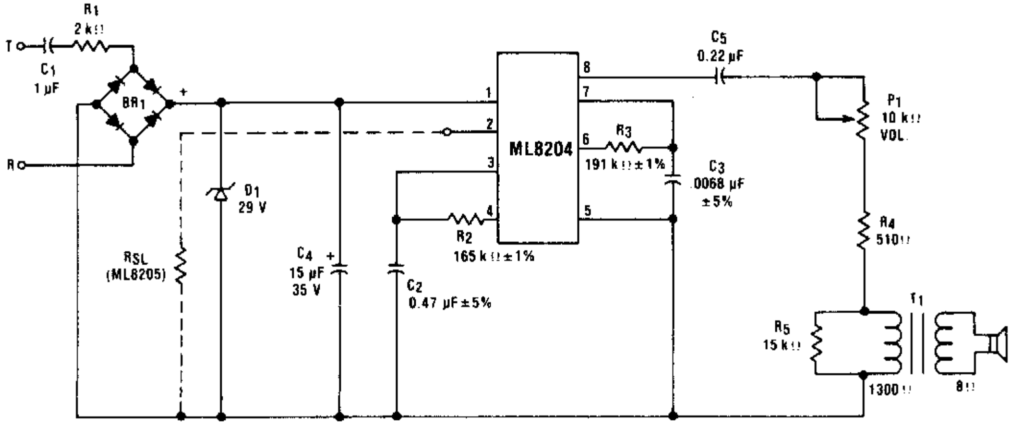

R1
2 kΩ
T
C1
1 μF
BR1
+
R
D1
29 V
RSL
(ML8205)
C4
+
15 μF
35 V
1
2
3
4
ML8204
R2
165 kΩ ± 1%
C2
0.47 μF ± 5%
8
7
6
5
R3
191 kΩ ± 1%
C3
.0068 μF
± 5%
C5
0.22 μF
P1
10 kΩ
VOL.
R4
510Ω
T1
R5
15 kΩ
1300Ω
8Ω

Fig. 85-5

Circuit Notes

This circuit uses ML8204/ML8205 devices. With the components shown, the output frequency chops between 512 Hz (f_{H1}) and 640 Hz (f_{H2}) at a 10 Hz (f_L) rate.

TELEPHONE LINE MONITOR

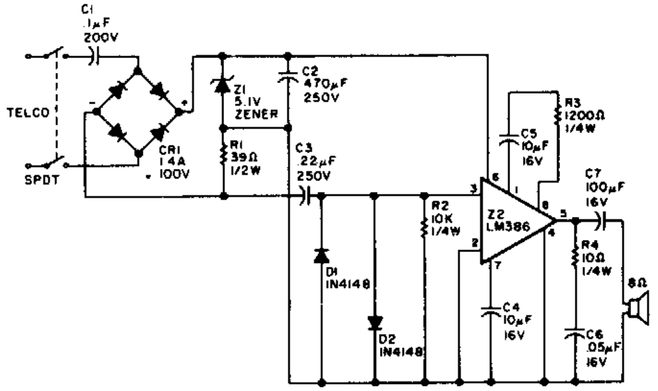

C1
.1μF
200V
TELCO
SPDT
-
+
CR1
1.4A
100V
Z1
5.1V
ZENER
R1
39Ω
1/2W
C2
470μF
250V
C3
.22μF
250V
D1
IN4148
D2
IN4148
R2
10K
1/4W
3
2
6
1
7
8
5
4
Z2
LM386
C5
10μF
16V
R3
1200Ω
1/4W
C4
10μF
16V
C7
100μF
16V
R4
10Ω
1/4W
C6
.05μF
16V
8Ω

Fig. 85-6

Circuit Notes

Using rectified audio as a power supply, this monitor will send the telephone line audio into an 8 ohm speaker.

TONE DIAL GENERATOR

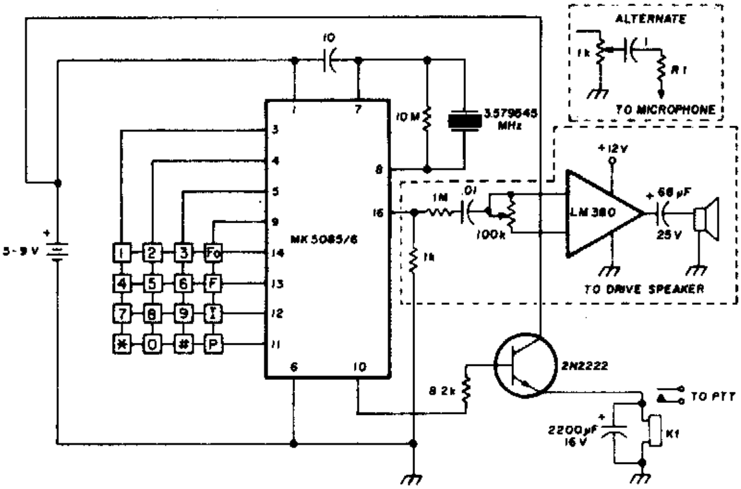

ALTERNATE
1k
1
R1
TO MICROPHONE
10
1
7
10 M
3.579545 MHz
3
4
5
9
8
16
14
13
12
11
6
10
MK 5085/6
5-9 V
+
1 2 3 Fo
4 5 6 F
7 8 9 I
* 0 # P
+12V
1M
.01
100k
1k
LM 380
+ 68 µF
25 V
TO DRIVE SPEAKER
2N2222
8.2k
TO PTT
2200µF 16 V
K1

Fig. 85-7

Circuit Notes

The circuit requires a minimum of parts and uses a low cost standard 3.579545-MHz television color-burst crystal. The speaker can be eliminated and the output fed directly into the microphone input of a transmitter.

TONE DIAL ENCODER

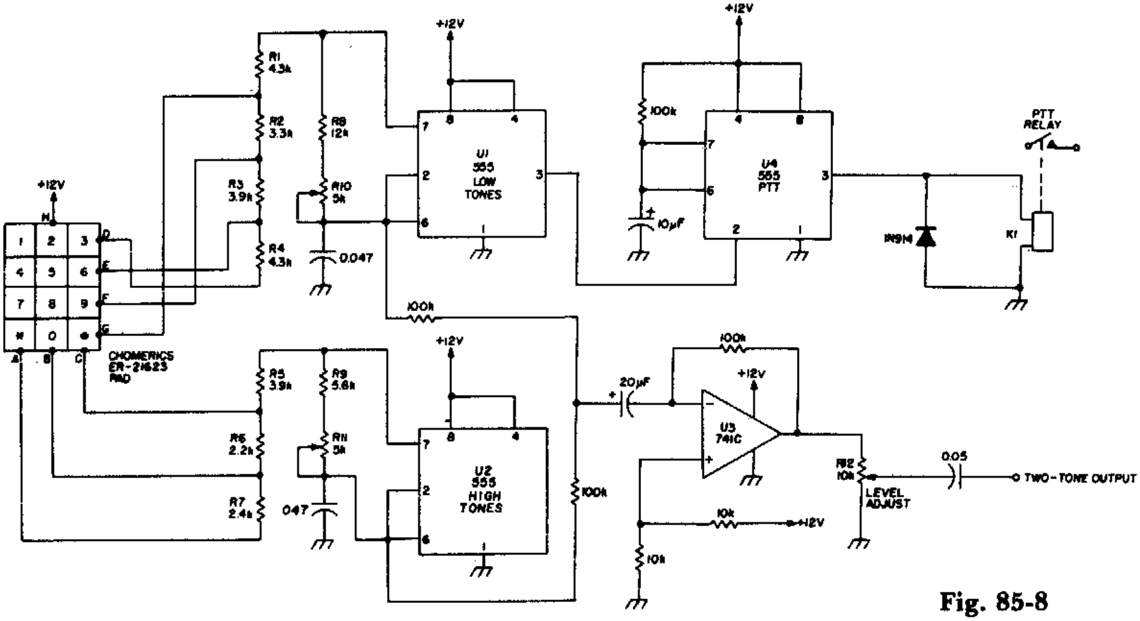
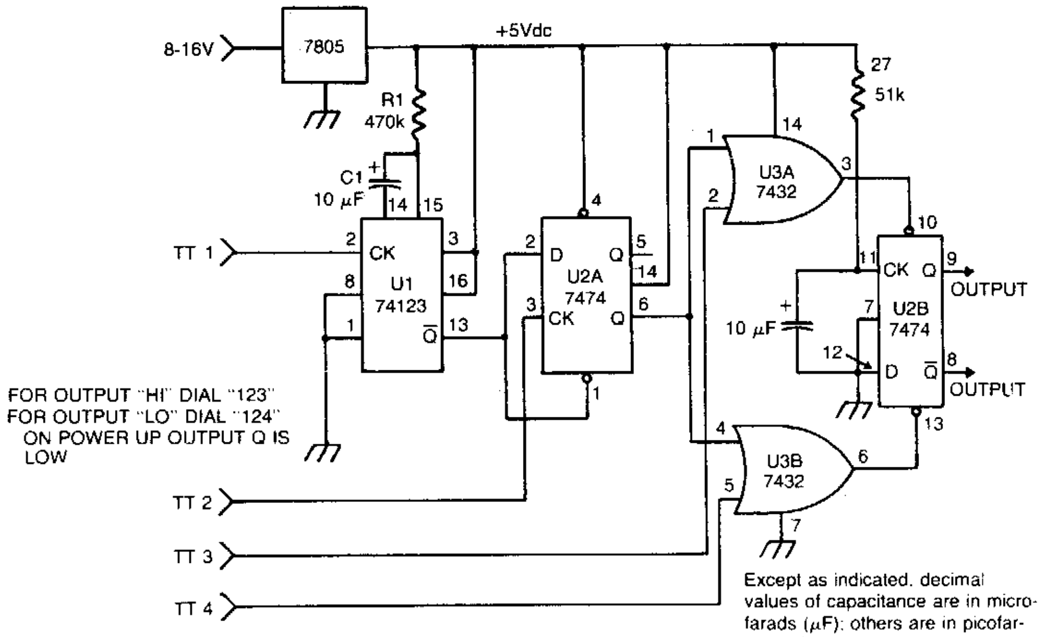

+12V
R1 4.3k
R2 3.3k
R3 3.9k
R4 4.3k
R8 12k
R10 5k
0.047
U1 555 LOW TONES
7
8
4
2
3
6
1
+12V
100k
10µF
U4 555 PTT
PTT RELAY
IN914
K1
+12V
H
1 2 3
4 5 6
7 8 9
* 0 #
D
E
F
G
A
B
C
CHOMERICS ER-21623 PAD
100k
R5 3.9k
R6 2.2k
R7 2.4k
R9 5.6k
R11 5k
.047
U2 555 HIGH TONES
100k
20µF
100k
U3 741C
R12 10k
LEVEL ADJUST
0.05
TWO-TONE OUTPUT
10k
+12V
10k

Fig. 85-8

Circuit Notes

Tone dial encoder with automatic PTT control uses the 555 timers.

TONE DIAL SEQUENCE DECODER


8-16V
7805
+5Vdc
R1
470k
C1
10 μF
TT 1
CK
U1
74123
Q̄
D
Q
U2A
7474
CK
Q
27
51k
U3A
7432
U3B
7432
CK Q
U2B
7474
D Q̄
OUTPUT
OUTPUT
10 μF
FOR OUTPUT "HI" DIAL "123"
FOR OUTPUT "LO" DIAL "124"
ON POWER UP OUTPUT Q IS LOW
TT 2
TT 3
TT 4

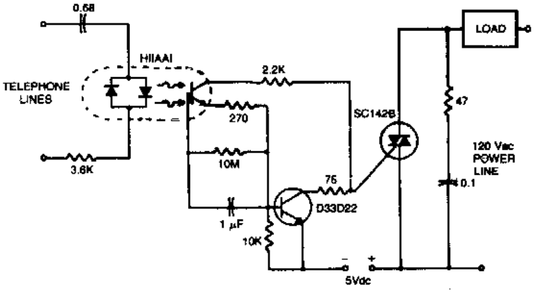
Except as indicated, decimal values of capacitance are in microfarads (μF); others are in picofarads (pF); resistances are in ohms. k = 1,000 M = 1,000,000

Fig. 85-9

Circuit Notes

The circuit takes active low inputs from a Touch Tone decoder and reacts to a proper sequence of digits. The proper sequence is determined by which Touch Tone digits the user connects to the sequence decoder inputs TT1, TT2, TT3, and TT4.

REMOTE RING EXTENDER SWITCH

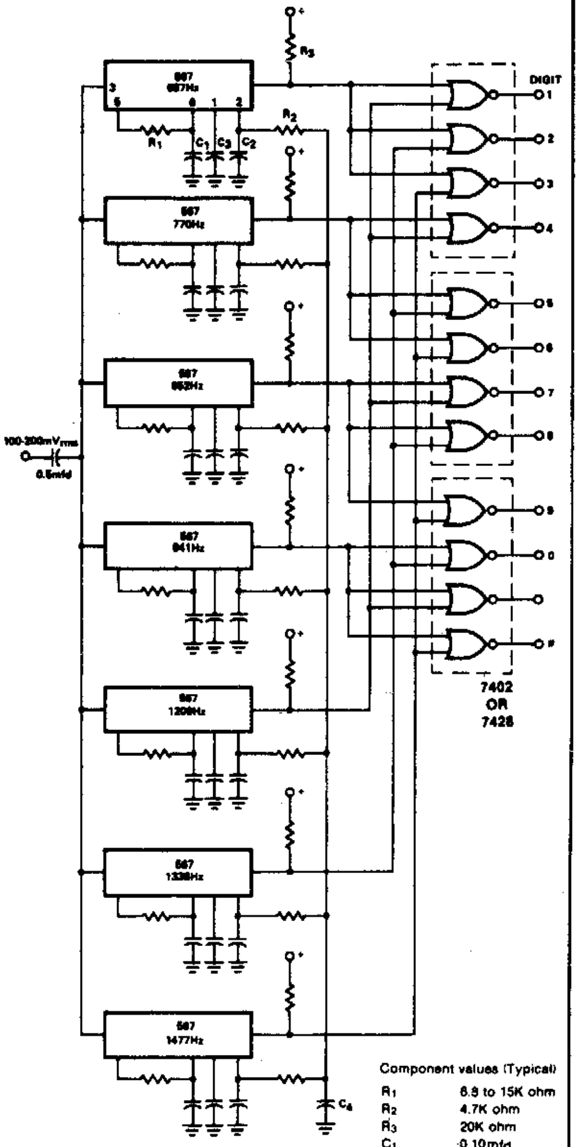

0.68
TELEPHONE LINES
HIIAAI
2.2K
270
10M
3.6K
1 μF
10K
75
D33D22
SC142B
47
LOAD
120 Vac POWER LINE
0.1
5Vdc

Circuit Notes

The circuit can operate lamps and buzzers from the 120 V, 60 Hz power line while maintaining positive isolation between the telephone line and the power line. Use of the isolated tab triac simplifies heat sinking by removing the constraint of isolating the triac heat sink from the chassis.

Fig. 85-10

TONE DIAL DECODER

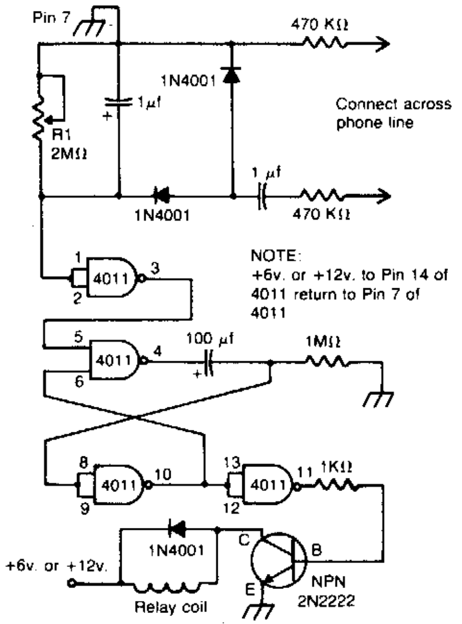

+
R3
567
697Hz
3
5
6
1
2
R1
C1
C3
C2
R2
DIGIT
1
2
3
4
567
770Hz
5
6
567
852Hz
7
8
100-200mV rms
0.5mfd
567
941Hz
9
0
#
567
1209Hz
7402
OR
7428
567
1336Hz
567
1477Hz
C4

Component values (Typical)

R_1	6.8 to 15K ohm
R_2	4.7K ohm
R_3	20K ohm
C_1	0.10mfd
C_2	1.0mfd 5V
C_3	2.2mfd 6V
C_4	250 μF 6V

Fig. 85-11

TELEPHONE RELAY


Pin 7
470 KΩ
1N4001
1 μf
+
Connect across phone line
R1
2MΩ
1 μf
1N4001
470 KΩ
1
3
4011
2
NOTE:
+6v. or +12v. to Pin 14 of 4011 return to Pin 7 of 4011
5
4011
4
100 μf
+
1MΩ
6
8
4011
10
9
13
4011
11
1KΩ
12
1N4001
C
B
+6v. or +12v.
E
NPN
2N2222
Relay coil

Circuit Notes

Connected across the bell circuit of phone, this circuit closes a relay when the phone is ringing. Use the delay contacts to actuate any bell, siren, buzzer or lamp.

Fig. 85-12

TELEPHONE-CONTROLLED TAPE STARTER (TCTS)

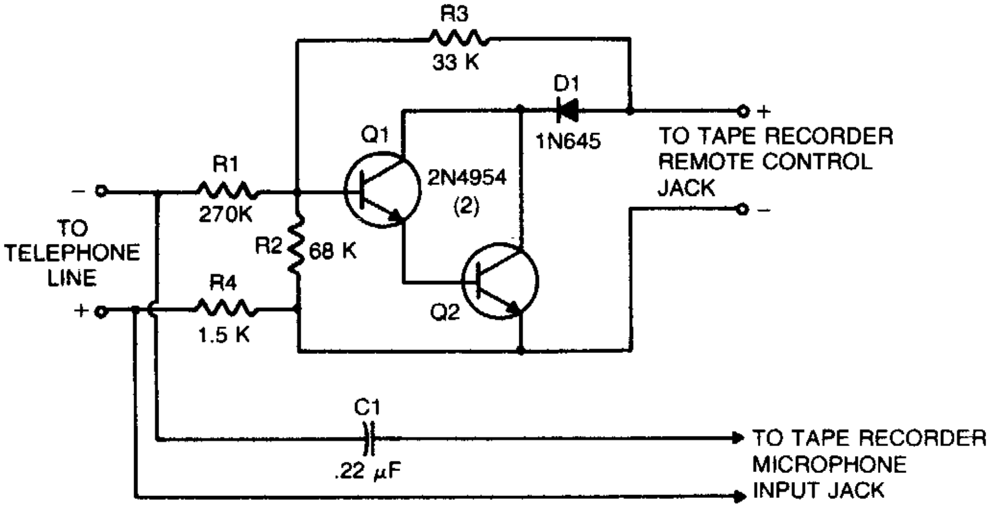

R3
33 K
D1
1N645
+
TO TAPE RECORDER
REMOTE CONTROL
JACK
−
Q1
2N4954
(2)
R1
270K
−
TO
TELEPHONE
LINE
+
R2
68 K
R4
1.5 K
Q2
C1
.22 μF
TO TAPE RECORDER
MICROPHONE
INPUT JACK

Fig. 85-13

Circuit Notes

This circuit converts a tape recorder into a completely automatic telephone conversation recording instrument that needs no external power source. Voltage at the switch terminals of tape recorder applied to a pair of Darlington-connected transistors, Q1 and Q2, will turn on and start the tape recorder. To turn the transistors off, and thereby stop the machine, apply a negative voltage to the base of Q1 from the phone line. When the telephone receiver is on the hook, there is typically about 50 volts dc across the phone divided across R1, R2, and R4 in such a way that the base of Q1 is sufficiently negative to keep the tape recorder off. When the phone's receiver is picked up, the voltage on the telephone line drops to about 5 volts, which leaves insufficient negative voltage on the base of Q1 to keep it cut off, so the tape recorder starts and begins to record.

TELEPHONE-LINE POWERED REPERTORY DIALER

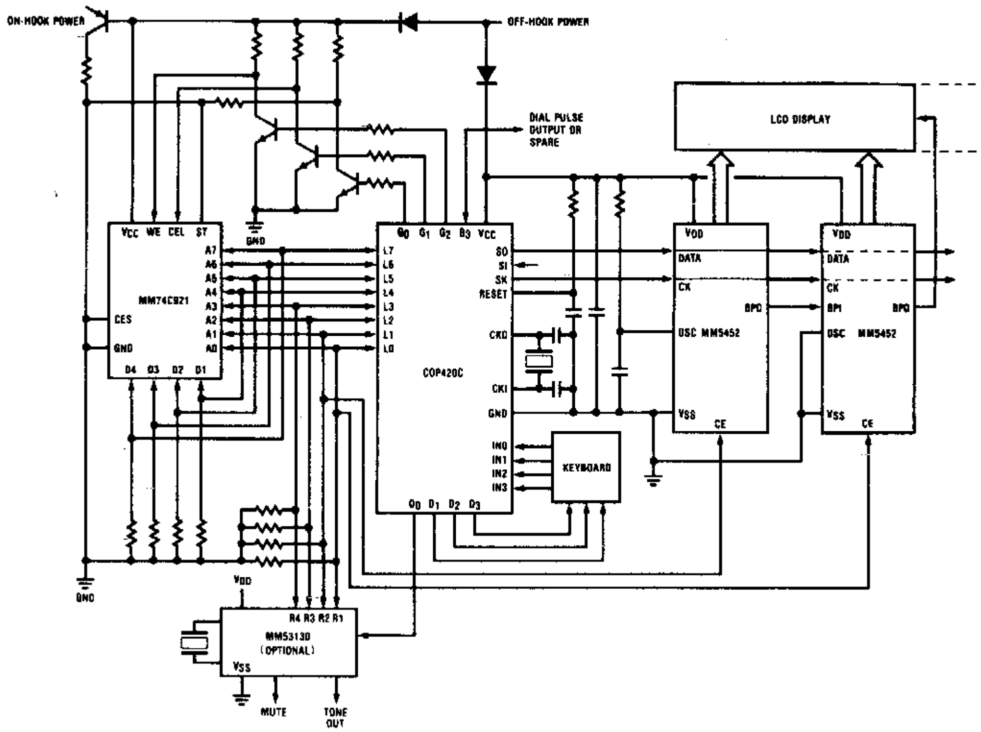

ON-HOOK POWER
OFF-HOOK POWER
DIAL PULSE OUTPUT OR SPARE
LCD DISPLAY
GND
Vcc
WE
CEL
ST
A7
A6
A5
A4
A3
A2
A1
A0
MM74C921
CES
GND
D4
D3
D2
D1
G0
G1
G2
G3
Vcc
L7
L6
L5
L4
L3
L2
L1
L0
SO
SI
SK
RESET
CKO
COP420C
CKI
GND
IN0
IN1
IN2
IN3
D0
D1
D2
D3
KEYBOARD
VDD
DATA
CK
BPO
OSC MM5452
VSS
CE
VDD
DATA
CK
BPI
BPO
OSC MM5452
VSS
CE
GND
VDD
R4 R3 R2 R1
MM53130
(OPTIONAL)
VSS
MUTE
TONE OUT

Fig. 85-14

Circuit Notes

Repertory dialer phone has a library of fifteen frequently used numbers, (plus the last number dialed) stored in a standard CMOS RAM. A pushbutton keyboard enables telephone numbers to be keyed in and dialed out directly or a telephone number to be stored in the RAM and dialed automatically.

TELEPHONE OFF-HOOK INDICATOR

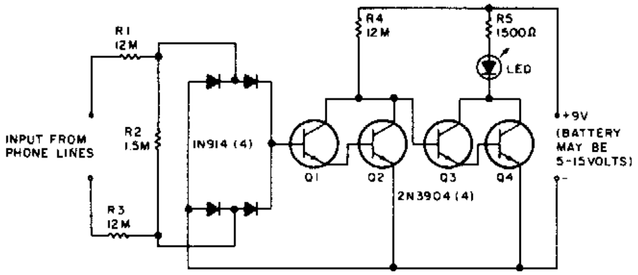

R1
12M
R2
1.5M
R3
12M
INPUT FROM PHONE LINES
IN914 (4)
R4
12M
R5
1500Ω
LED
Q1
Q2
Q3
Q4
2N3904 (4)
+9V
(BATTERY MAY BE 5-15VOLTS)
-

Fig. 85-15

Circuit Notes

The LED flickers when the phone is ringing or being dialed. It glows steadily when the phone is off the hook.

TELEPHONE HANDSET TONE DIAL ENCODER

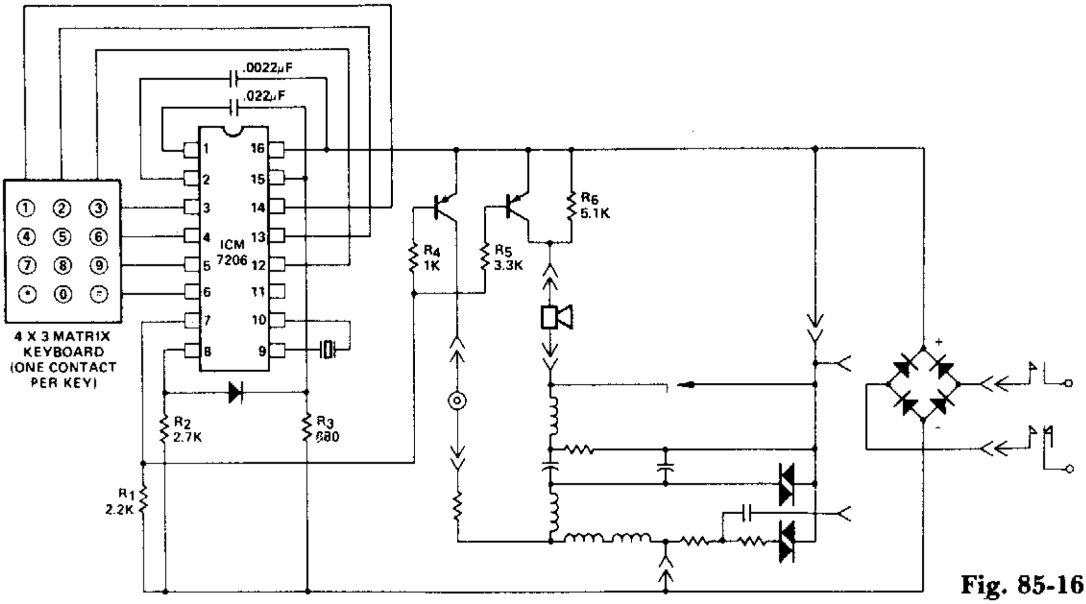

.0022µF
.022µF
ICM 7206
1
2
3
4
5
6
7
8
16
15
14
13
12
11
10
9
R6 5.1K
R4 1K
R5 3.3K
R2 2.7K
R3 680
R1 2.2K
4 X 3 MATRIX KEYBOARD (ONE CONTACT PER KEY)
+
-

Fig. 85-16

Circuit Notes

This encoder uses a single contact per key keyboard and provides all other switching function electronically. The diode between terminals 8 and 15 prevents the output going more than 1 volt negative with respect to the negative supply V−. The circuit operates over the supply voltage range from 3.5 volts to 15 volts.

LOW LINE LOADING RING DETECTOR

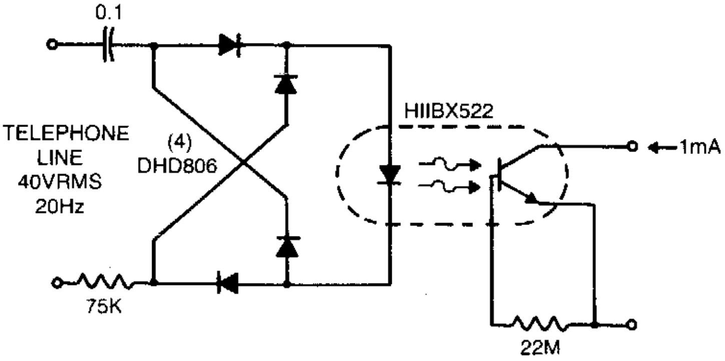
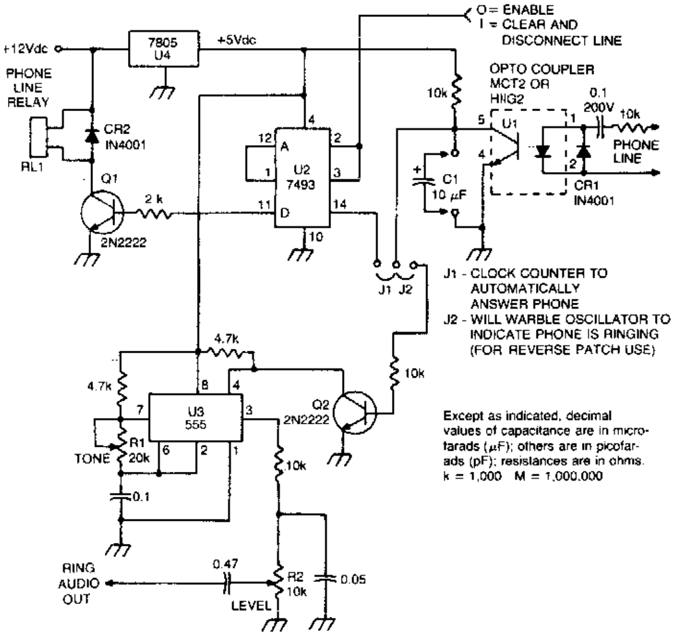

0.1
TELEPHONE LINE 40VRMS 20Hz
(4) DHD806
HIIBX522
←1mA
75K
22M

Fig. 85-17

Circuit Notes

Low line current loading is provided by the H11BX522 photodarlington optocoupler, which provides a 1 mA output from a 0.5 mA input.

PHONE AUTO ANSWER AND RING INDICATOR


+12Vdc
7805
U4
+5Vdc
PHONE
LINE
RELAY
RL1
CR2
IN4001
Q1
2N2222
2 k
12
A
U2
7493
1
11
D
4
2
3
14
10
O = ENABLE
I = CLEAR AND
DISCONNECT LINE
OPTO COUPLER
MCT2 OR
H11G2
10k
5
U1
4
1
2
0.1
200V
10k
PHONE
LINE
CR1
IN4001
C1
10 µF
J1 J2
4.7k
4.7k
10k
8
4
7
U3
555
3
Q2
2N2222
R1
TONE
20k
6
2
1
0.1
10k
RING
AUDIO
OUT
0.47
R2
10k
LEVEL
0.05

J1 - CLOCK COUNTER TO AUTOMATICALLY ANSWER PHONE
J2 - WILL WARBLE OSCILLATOR TO INDICATE PHONE IS RINGING (FOR REVERSE PATCH USE)

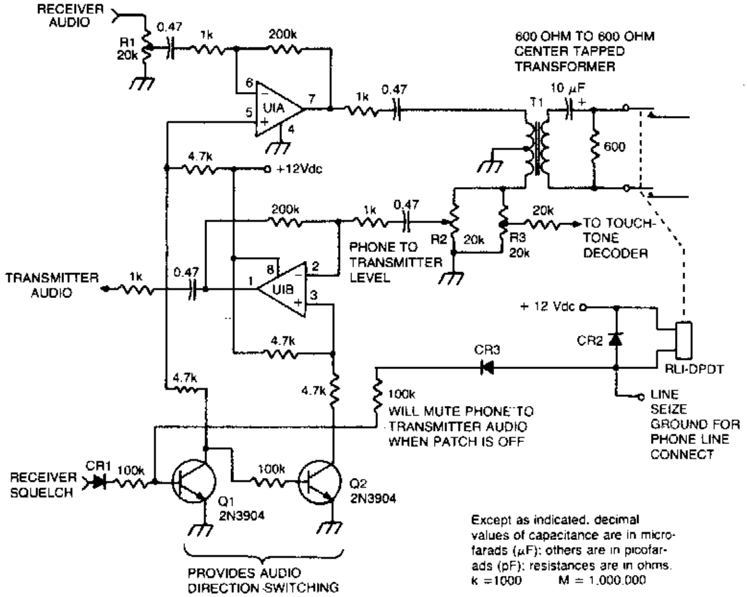
Except as indicated, decimal values of capacitance are in microfarads (µF); others are in picofarads (pF); resistances are in ohms.
k = 1,000 M = 1,000,000

Circuit Notes

Ring detect circuit for automatic phone answering or tone generation for reverse autopatch use.

Fig. 85-18

AUTOPATCH TELEPHONE LINE INTERFACE


RECEIVER
AUDIO
R1
20k
0.47
1k
200k
6
5
UIA
7
4
1k
0.47
600 OHM TO 600 OHM
CENTER TAPPED
TRANSFORMER
10 µF
T1
600
4.7k
+12Vdc
200k
1k
0.47
R2
20k
R3
20k
20k
TO TOUCH-
TONE
DECODER
PHONE TO
TRANSMITTER
LEVEL
TRANSMITTER
AUDIO
1k
0.47
8
1
UIB
2
3
+ 12 Vdc
CR2
RLI-DPDT
4.7k
CR3
4.7k
4.7k
100k
WILL MUTE PHONE TO
TRANSMITTER AUDIO
WHEN PATCH IS OFF
LINE
SEIZE
GROUND FOR
PHONE LINE
CONNECT
RECEIVER
SQUELCH
CR1
100k
Q1
2N3904
100k
Q2
2N3904
PROVIDES AUDIO
DIRECTION SWITCHING

Except as indicated, decimal values of capacitance are in microfarads (µF); others are in picofarads (pF); resistances are in ohms.
k = 1000 M = 1,000,000

Circuit Notes

This circuit provides for the receiver-to-phone line and phone line-to-transmitter link, with both using an op amp for gain.

Fig. 85-19

TELEPHONE RINGER USES PIEZOELECTRIC DEVICE

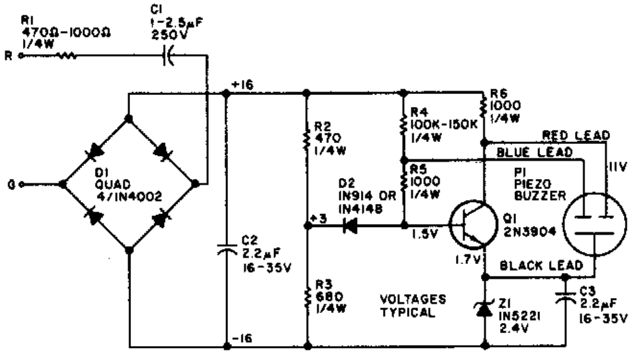

R1
470Ω-1000Ω
1/4W
C1
1-2.5μF
250V
R
G
+16
-16
D1
QUAD
4/1N4002
C2
2.2μF
16-35V
R2
470
1/4W
R3
680
1/4W
+3
D2
1N914 OR
1N4148
R4
100K-150K
1/4W
R5
1000
1/4W
R6
1000
1/4W
1.5V
1.7V
Q1
2N3904
RED LEAD
BLUE LEAD
BLACK LEAD
11V
P1
PIEZO
BUZZER
VOLTAGES
TYPICAL
Z1
1N5221
2.4V
C3
2.2μF
16-35V

Fig. 85-20

Circait Notes

The electronic bell needs no power supply. Most of the resistors are not critical, although C2, R2, and R3 work best at the values given. Leaving out R1 will make the unit ring louder. The piezo buzzer may vary from store to store. If it has two leads, connect the red lead to the collector and the black lead to the emitter of Q1. If a third (blue) lead is present, connect it to the base of Q1.

ELECTRONIC PHONE BELL

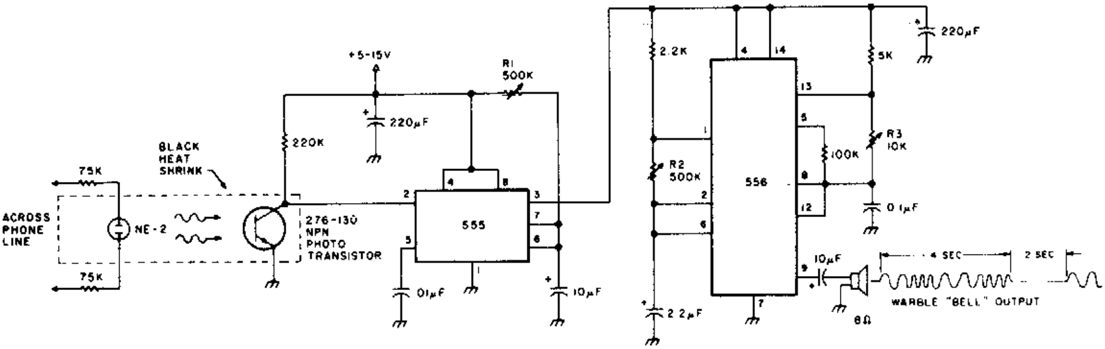

+5-15V
R1
500K
220μF
220K
BLACK
HEAT
SHRINK
75K
75K
ACROSS
PHONE
LINE
NE-2
276-130
NPN
PHOTO
TRANSISTOR
555
.01μF
10μF
2.2K
R2
500K
2.2μF
556
5K
220μF
R3
10K
100K
0.1μF
10μF
8Ω
4 SEC
2 SEC
WARBLE "BELL" OUTPUT

Fig. 85-21

Circuit Notes

The speaker emits a distinctive warble tone when ring pulses are applied to the phone line. Use this circuit as a remote bell or disconnect the phone's ringer for direct use. R1 adjusts the duration of the output; R2 and R3 control the tone's duty cycle and frequency. The transistor is a general-purpose NPN photodevice. The neon bulb and transistor are coupled with the heat-shrink tubing to form an optoisolator.

86

Temperature Controls

The sources of the following circuits are contained in the Sources section beginning on page 730. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

Boiler Control
Heater Control
Two-Wire Remote AC Electronic Thermostat
Three-Wire Electronic Thermostat
Temperature-Sensitive Heater Control
Temperature Controller
Single-Setpoint Temperature Controller
Temperature Controller
Temperature Control
Temperature Controller
Temperature Controller
Portable Calibrator

BOILER CONTROL

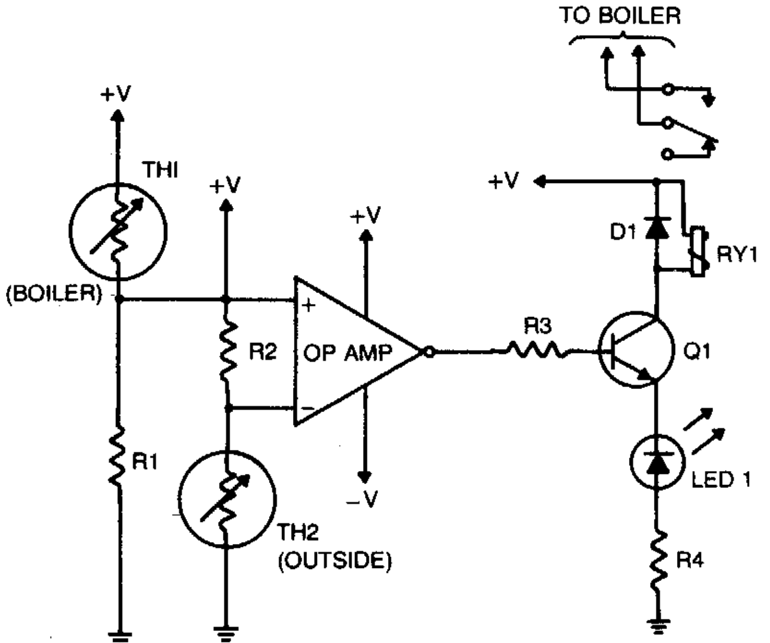
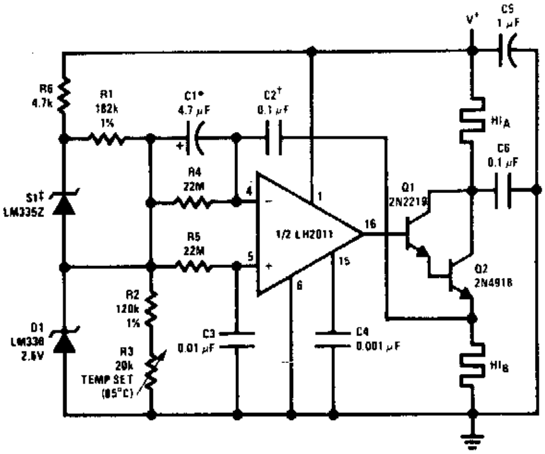


Fig. 86-1

Circuit Notes

The purpose of this circuit is to control the water temperature in a hot-water heating system. What it does is to lower the boiler temperature as the outside air temperature increases. The op amp is used as a comparator. Thermistor TH2 and R2 form a voltage divider that supplies a reference voltage to the op-amp's inverting input. Thermistor TH2 is placed outdoors, and the values of TH2 and R2 should be chosen so that when the outside temperature is 25 °F, the resistance of the thermistor and resistor are equal. Resistor R1 and thermistor TH1 make up a voltage divider that supplies a voltage to the op amp's noninverting input. Thermistor TH1 is placed inside the boiler and the values of TH1 and R1 should be chosen so that when the boiler's temperature is 160 °F, their resistances are equal. The output of the op amp controls Q1, which is configured as a transistor switch. When the logic output of the op amp is high, Q1 is turned on, energizing relay RY1. The relay's contacts should be wired so that the boiler's heat supply is turned off (relay energized).

HEATER CONTROL


V+
C5
1 μF
R6
4.7k
R1
182k
1%
C1*
4.7 μF
C2†
0.1 μF
HIA
C6
0.1 μF
S1‡
LM335Z
R4
22M
4
1
Q1
2N2219
R5
22M
1/2 LH2011
16
5
15
6
Q2
2N4918
R2
120k
1%
D1
LM336
2.5V
C3
0.01 μF
C4
0.001 μF
R3
20k
TEMP SET
(85°C)
HIB

* solid tantalum

† mylar

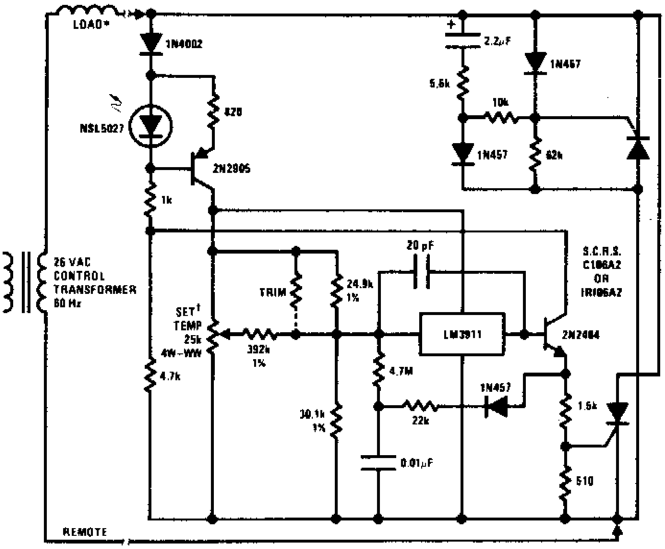
‡ close thermal coupling between sensor and oven shell is recommended.

Circuit Notes

This proportional control crystal oven heater uses lead/lag compensation for fast setting. The time constant is changed with R4 and compensating resistor R5. If Q2 is inside the oven, a regulated supply is recommended for 0.1 °C. control.

Fig. 86-2

TWO-WIRE REMOTE AC ELECTRONIC THERMOSTAT (GAS OR OIL FURNACE CONTROL)


LOAD*
1N4002
2.2μF
1N457
5.6k
10k
820
NSL5027
1N457
82k
2N2905
1k
20 pF
S.C.R.S.
C106A2
OR
IR106A2
26 VAC
CONTROL
TRANSFORMER
60 Hz
TRIM
24.9k
1%
SET†
TEMP
25k
4W–WW
392k
1%
LM3911
2N2484
4.7k
4.7M
1N457
1.5k
30.1k
1%
22k
0.01μF
510
REMOTE

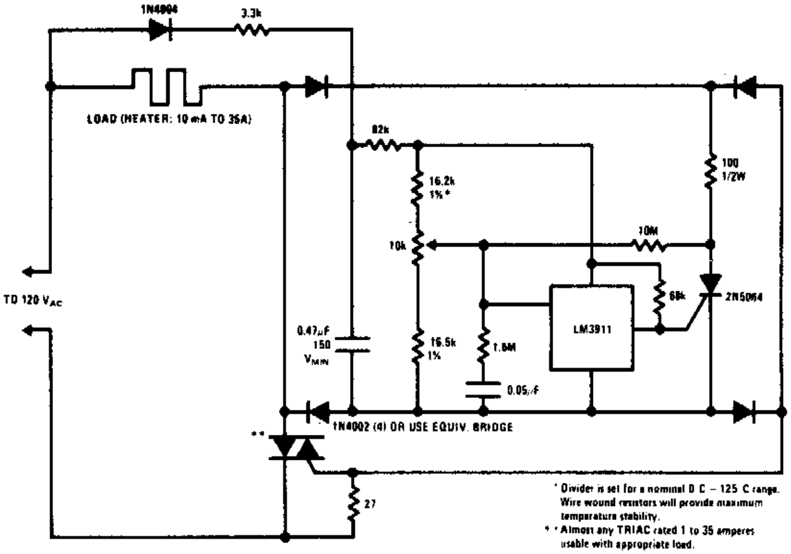
* Solenoid or 6–15W heater

† Pot will provide about a 50°F to 90°F setting range. The trim resistor (100k) is selected to bring 70°F near the middle of the pot rotation.

SCR heating, by proper positioning, can preheat the sensor giving control anticipation as is presently used in many home thermostats.

Fig. 86-3

THREE-WIRE ELECTRONIC THERMOSTAT


1N4004
3.3k
LOAD (HEATER: 10 mA TO 35A)
82k
16.2k
1%*
10k
10M
100
1/2W
68k
2N5064
LM3911
0.47μF
150
V_MIN
19.5k
1%
1.6M
0.05μF
1N4002 (4) OR USE EQUIV. BRIDGE
**
27
TO 120 V_AC

* Divider is set for a nominal 0 C – 125 C range. Wire wound resistors will provide maximum temperature stability.

** Almost any TRIAC rated 1 to 35 amperes usable with appropriate load.

Fig. 86-4

TEMPERATURE-SENSITIVE HEATER CONTROL

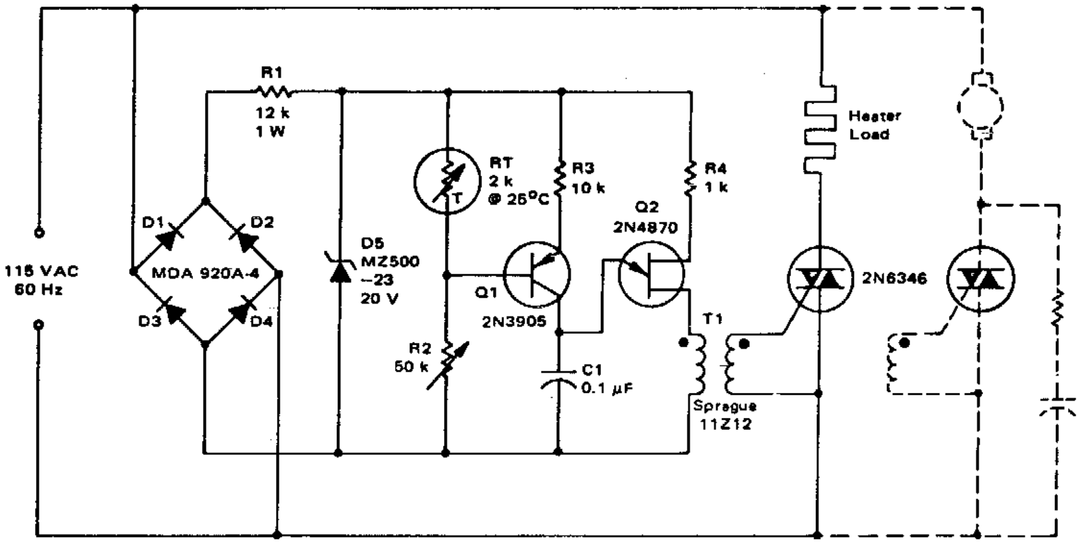
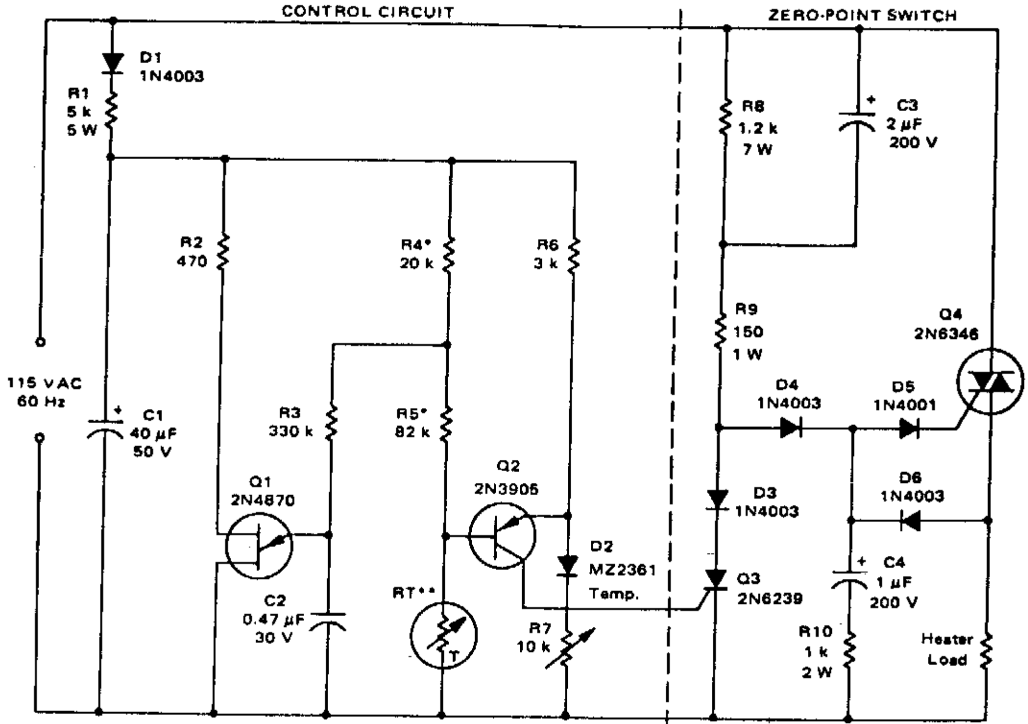

R1
12 k
1 W
115 VAC
60 Hz
D1
D2
MDA 920A-4
D3
D4
D5
MZ500
–23
20 V
RT
2 k
@ 25°C
T
R3
10 k
R4
1 k
Q2
2N4870
Q1
2N3905
R2
50 k
C1
0.1 μF
T1
Sprague
11Z12
Heater
Load
2N6346

Fig. 86-5

TEMPERATURE CONTROLLER


CONTROL CIRCUIT
ZERO-POINT SWITCH
D1
1N4003
R1
5 k
5 W
R2
470
R4*
20 k
R6
3 k
R8
1.2 k
7 W
C3
2 μF
200 V
R9
150
1 W
Q4
2N6346
115 VAC
60 Hz
C1
40 μF
50 V
R3
330 k
R5*
82 k
D4
1N4003
D5
1N4001
Q1
2N4870
Q2
2N3905
D3
1N4003
D6
1N4003
D2
MZ2361
Temp.
Q3
2N6239
C4
1 μF
200 V
C2
0.47 μF
30 V
RT**
T
R7
10 k
R10
1 k
2 W
Heater
Load

*Low Temp. Coefficient
**Fenwell QR51J1 100 k Thermistor

Fig. 86-6

SINGLE-SETPOINT TEMPERATRE CONTROLLER

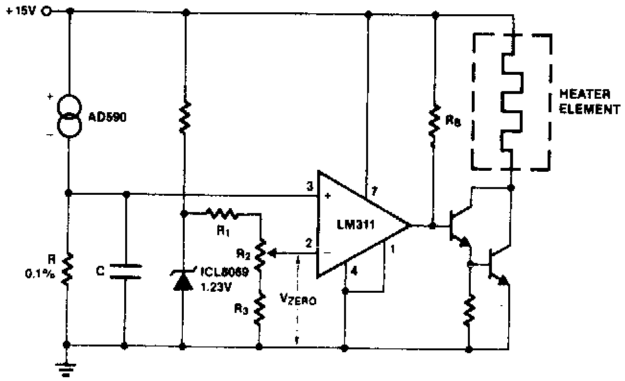
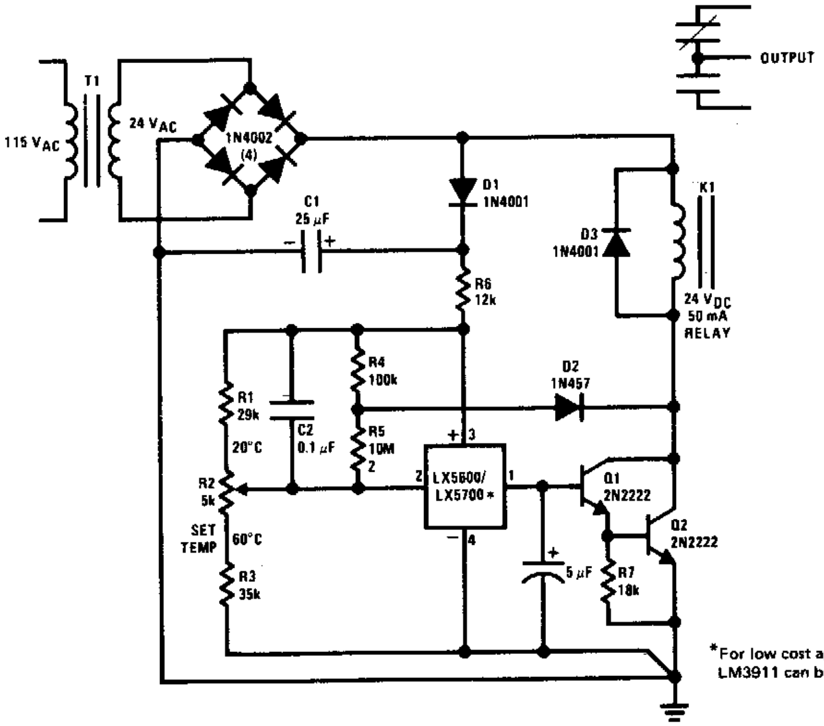

+15V
AD590
+
−
HEATER ELEMENT
R₆
3
7
+
LM311
R₁
2
1
R₂
−
4
R
0.1%
C
ICL8069
1.23V
R₃
V_ZERO

Fig. 86-7

Circuit Notes

The AD590 produces a temperature-dependent voltage across R (C is for filtering noise). Setting R2 produces a scale-zero voltage. For the Celsius scale, make R = 1 K and V_{zero} = 0.273 volts. For Fahrenheit, R = 1.8 K and V_{zero} = 0.460 volts.

TEMPERATURE CONTROLLER


OUTPUT
T1
115 VAC
24 VAC
1N4002 (4)
C1 25 µF
D1 1N4001
D3 1N4001
K1
24 VDC 50 mA RELAY
R6 12k
R4 100k
D2 1N457
R1 29k
C2 0.1 µF
R5 10M
20°C
R2 5k
SET TEMP
60°C
R3 35k
LX5600/ LX5700*
Q1 2N2222
Q2 2N2222
5 µF
R7 18k
1
2
3
4

*For low cost applications, an LM3911 can be used

Fig. 86-8

Circuit Notes

The sensor is a standard TO-5 or TO-46 package. For surface or air temperature sensing. Small clip-on heat sinks can be used. A simple probe can be made using heat-shrink tubing and RTV silicon rubber. Three-leads-plus-shield cable is a good choice for wire with the shield connected to pin 4. The controller can be used for baths, ovens, oven-temperature protection, or even home thermostats. Long-term stability and repeatability is better than 0.5 °C.

TEMPERATURE CONTROL

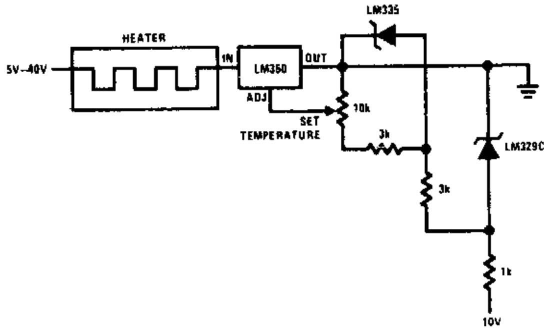
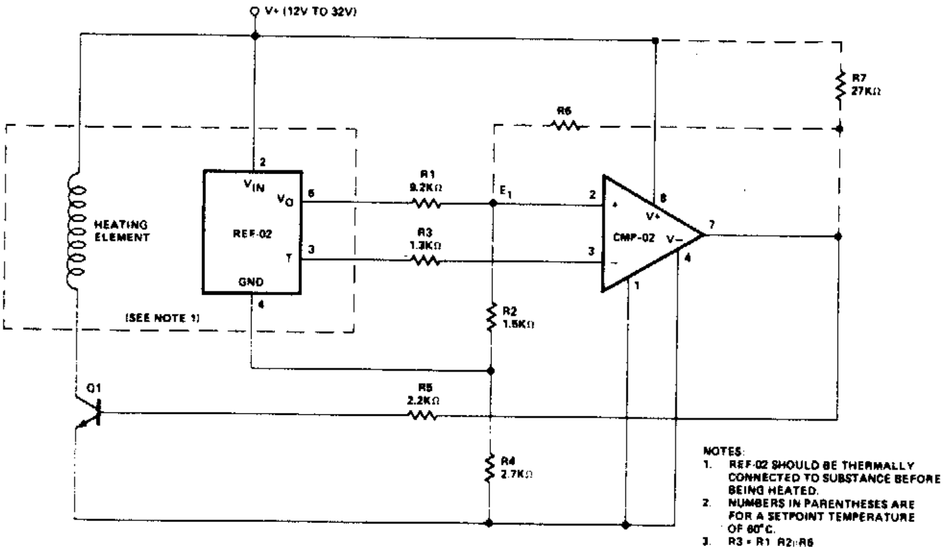

LM335
HEATER
5V–40V
IN
LM350
OUT
ADJ
10k
SET TEMPERATURE
3k
LM329C
3k
1k
10V

Fig. 86-9

TEMPEATURE CONTROLLER


V+ (12V TO 32V)
R7
27KΩ
R6
R1
9.2KΩ
R3
1.3KΩ
E1
HEATING ELEMENT
VIN
VO
T
GND
REF-02
CMP-02
V+
V−
(SEE NOTE 1)
R2
1.5KΩ
Q1
R5
2.2KΩ
R4
2.7KΩ
2
6
3
4
2
3
8
7
4
1

NOTES:
1. REF-02 SHOULD BE THERMALLY CONNECTED TO SUBSTANCE BEFORE BEING HEATED.
2. NUMBERS IN PARENTHESES ARE FOR A SETPOINT TEMPERATURE OF 60°C.
3. R3 = R1 R2‖R6

Fig. 86-10

Circuit Notes

Temperature control is achieved using the REF-02 +5 V Reference/Thermometer and a CMP-02 Precision Low Input Current Comparator. The CMP-02 turns on a heating element driver (Q1) whenever the present temperature drops below a setpoint temperature determined by the ratio of R1 to R2. The circuit also provides adjustable hysteresis and single supply operation.

TEMPERATURE CONTROLLER

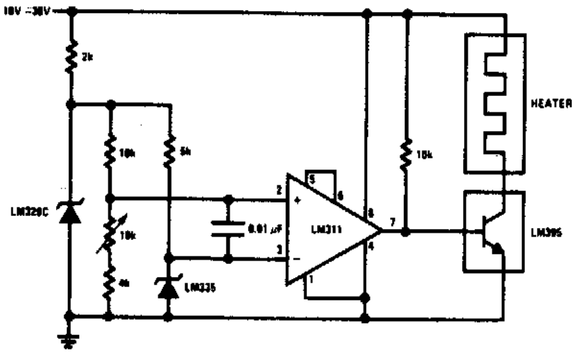
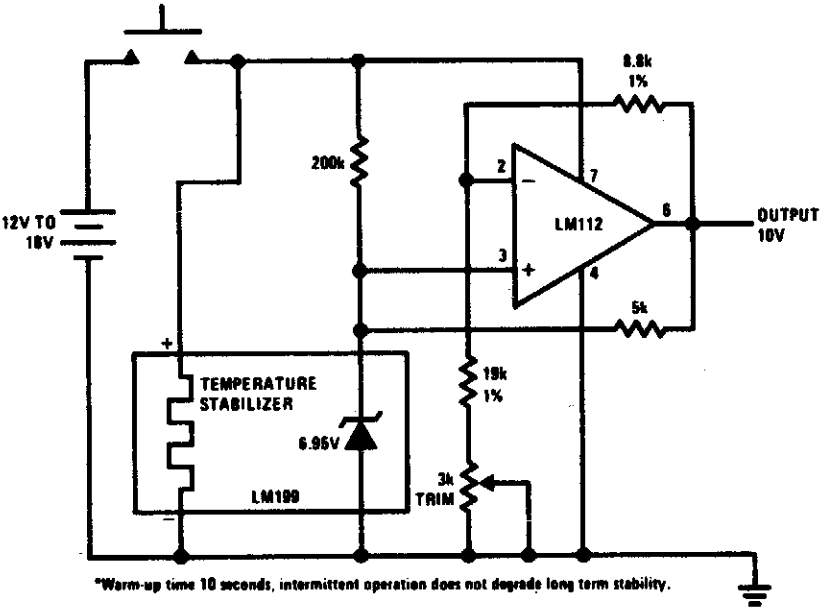

10V – 30V
2k
10k
5k
LM329C
10k
4k
LM335
0.01 μF
LM311
2
3
5
6
8
7
4
1
10k
HEATER
LM395

Fig. 86-11

PORTABLE CALIBRATOR


12V TO
18V
200k
8.8k
1%
2
3
7
6
4
LM112
OUTPUT
10V
5k
+
−
TEMPERATURE
STABILIZER
6.95V
LM199
19k
1%
3k
TRIM

*Warm-up time 10 seconds, intermittent operation does not degrade long term stability.

Fig. 86-12

87

Temperature Sensors

The sources of the following circuits are contained in the Sources section beginning on page 730. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

Linear Temperature-to-Frequency Transconducer
Temperature Meter
Four-Channel Temperature Sensor
Temperature Sensor
Integrated Circuit Temperature Sensor
Precision Temperature Transducer with Remote Sensor
Centigrade Calibrated Thermocouple Thermometer
μP Controlled Digital Thermometer
Isolated Temperature Sensor
Digital Thermometer
Variable Offset Thermometer
Differential Thermometer
Basic Digital Thermometer, Kelvin Scale
Basic Digital Thermometer, Kelvin Scale with Zero Adjust
Thermocouple Amplifier
Optical Pyrometer
Remote Temperature Sensing
Simple Differential Temperature Sensor
Differential Temperature Sensor
Centigrade Thermometer
Meter Thermometer with Trimmed Output
Kelvin Thermometer with Ground Referred Output
Lower Power Thermometer
0 °F-50 °F Thermometer
Temperature-to-Frequency Converter
0 °C-100 °C Thermometer
Ground Referred Fahrenheit Thermometer
Ground Referred Centigrade Thermometer
Ground Referred Centigrade Thermometer
Temperature Sensor
Positive Temperature Coefficient Resistor Temperature Sensor
Basic Digital Thermometer
Fahrenheit Thermometer

LINEAR TEMPERATURE-TO-FREQUENCY TRANSCONDUCER

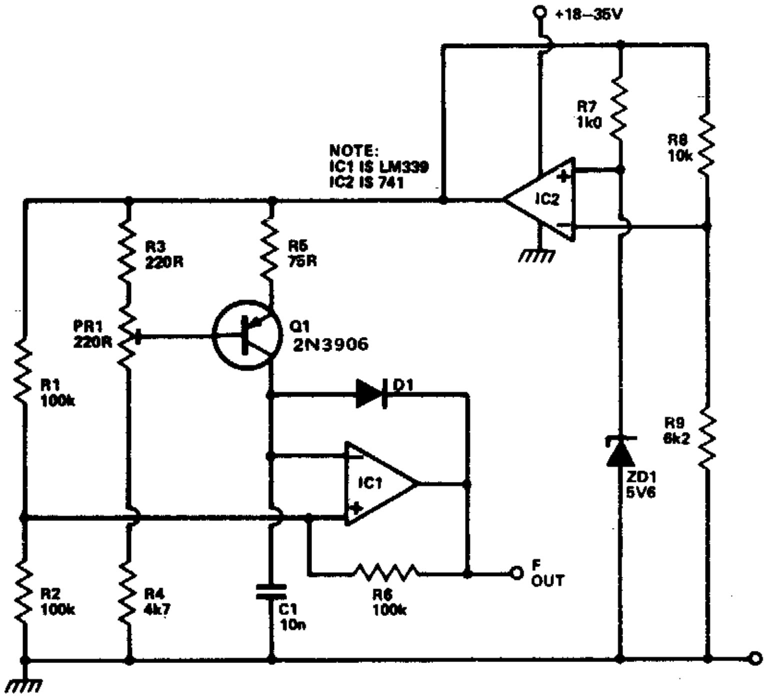

+18–35V
R7
1k0
R8
10k
NOTE:
IC1 IS LM339
IC2 IS 741
IC2
R3
220R
R5
75R
PR1
220R
Q1
2N3906
D1
R1
100k
IC1
R9
6k2
ZD1
5V6
F
OUT
R2
100k
R4
4k7
C1
10n
R6
100k

Fig. 87-1

Circuit Notes

This circuit provides a linear increase of frequency of 10 Hz/°C over 0-100 °C and can thus be used with logic systems, including microprocessors. Temperature probes Q1 V_{be} changes 2.2 mV/°C. This transistor is incorporated in a constant current source circuit. Thus, a current proportional to temperature will be available to charge C1. The circuit is powered via the temperature stable reference voltage supplied by the 741. Comparator IC1 is used as a Schmitt trigger whose output is used to discharge C1 via D1. To calibrate the circuit Q1 is immersed in boiling distilled water and PR1 adjusted to give 1 kHz output. The prototype was found to be accurate to within 0.2 °C.

TEMPERATURE METER

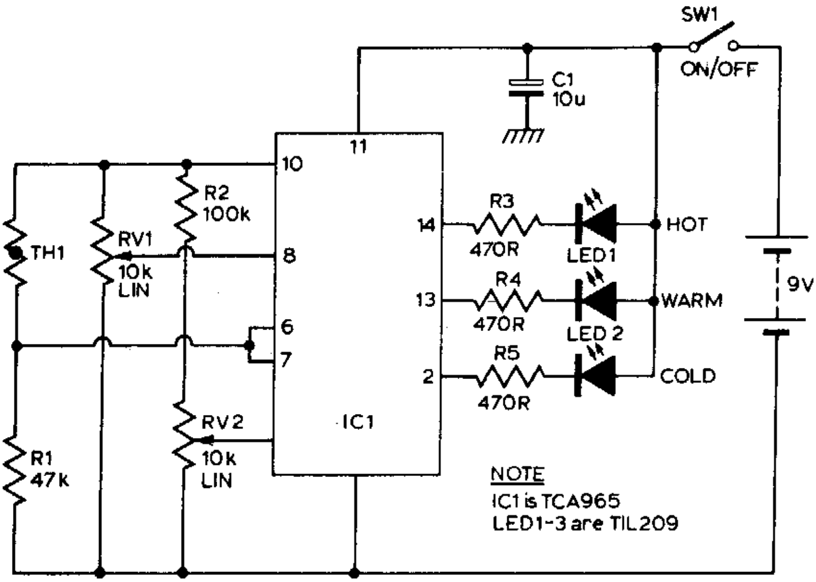

SW1
ON/OFF
C1
10u
11
10
R2
100k
R3
14
HOT
TH1
RV1
10k
LIN
8
470R
LED1
R4
13
WARM
9V
6
470R
LED 2
7
R5
2
COLD
470R
RV2
IC1
R1
47k
10k
LIN
NOTE
IC1 is TCA965
LED1-3 are TIL209

Fig. 87-2

Circuit Notes

TCA965 window discriminator IC allows the potentiometers RV1 and RV2 to set up a window height and window width respectively. R1 and thermistor TH1 for a potential divider connected across the supply lines. R1 is chosen such that at ambient temperature the voltage at the junction of these two components will be approximately half supply. As the temperature of the sensor changes, the voltage will change. RV1 will set the point which corresponds to the center voltage of a window the width of which is set by RV2. The switching points of the IC feature a Schmitt characteristic with low hysteresis. The outputs of IC1 indicate whether the input voltage is within the window or outside by virtue of being either too high or too low. The outputs of IC1 drive the LEDs via a current limiting resistor.

FOUR-CHANNEL TEMPERATURE SENSOR (0-50 °C)

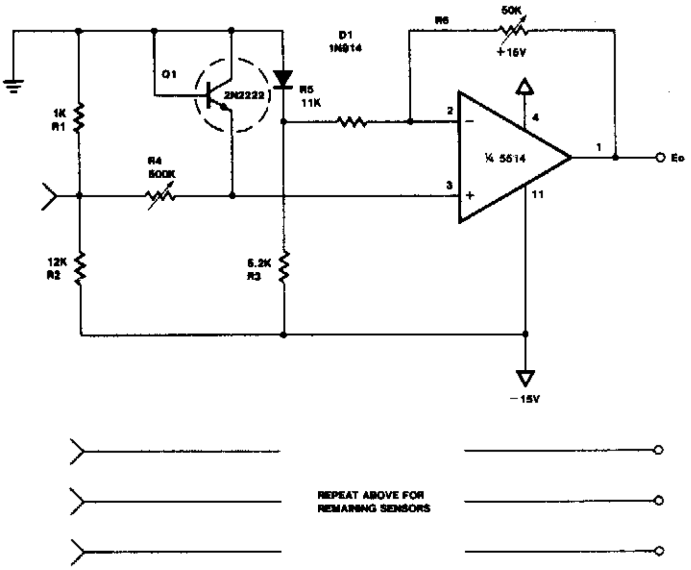
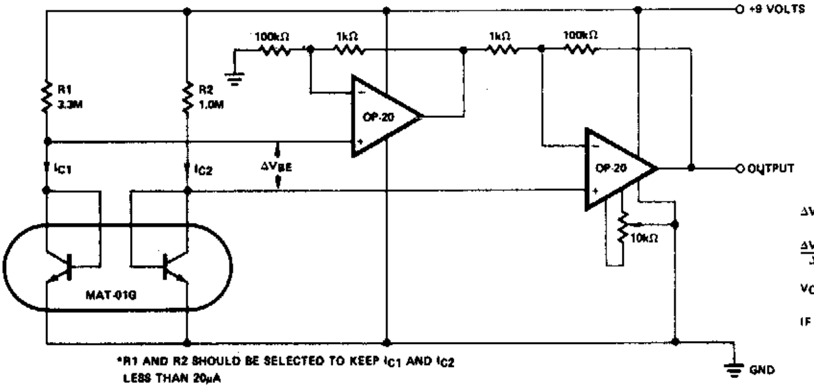

R6
50K
+15V
D1
1N914
Q1
2N2222
R5
11K
1K
R1
R4
500K
12K
R2
5.2K
R3
2
3
4
1
11
¼ 5514
Eo
−15V
REPEAT ABOVE FOR
REMAINING SENSORS

Fig. 87-3

TEMPERATURE SENSOR


+9 VOLTS
100kΩ
1kΩ
1kΩ
100kΩ
R1
3.3M
R2
1.0M
OP-20
OP-20
I_{C1}
I_{C2}
ΔV_{BE}
OUTPUT
10kΩ
MAT-01G
GND

*R1 AND R2 SHOULD BE SELECTED TO KEEP I_{C1} AND I_{C2} LESS THAN 20µA

DESIGN EQUATIONS

$$\Delta V_{BE} = \frac{kT}{q} \text{ LN } \left(\frac{I_{C2}}{I_{C1}}\right)$$

$$\frac{\Delta V_{BE}}{\Delta T} = 85.8 \text{ LN } \left(\frac{I_{C2}}{I_{C1}}\right) \ (\mu V/°K)$$

$$V_{OUT} = 101(\Delta V_{BE})$$

$$\text{IF } \frac{R1}{R2} = 3.2 \text{ THEN } TCV_{OUT} = 10mV/°C$$

Fig. 87-4

INTEGRATED CIRCUIT TEMPERATURE SENSOR

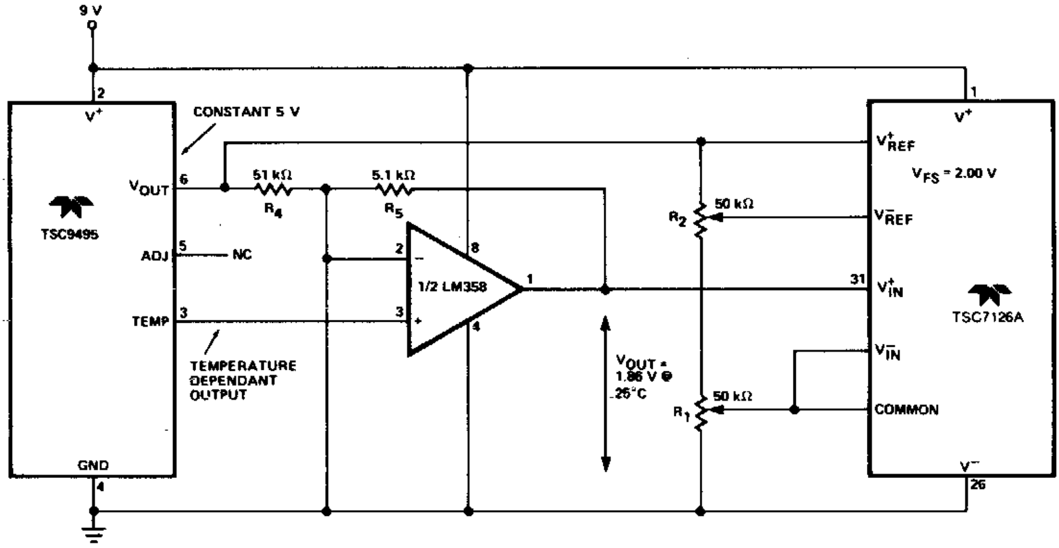
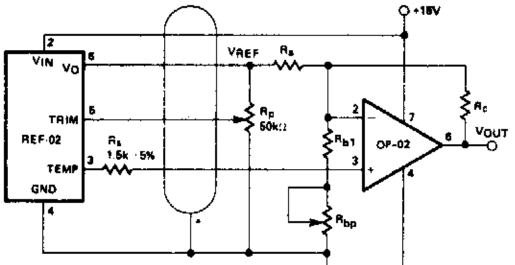

9 V
2
V+
CONSTANT 5 V
TSC9495
VOUT
6
ADJ
5
NC
TEMP
3
TEMPERATURE DEPENDANT OUTPUT
GND
4
51 kΩ
R4
5.1 kΩ
R5
2
−
3
+
1/2 LM358
8
4
1
VOUT = 1.86 V @ 25°C
R2
50 kΩ
R1
50 kΩ
1
V+
V+REF
VFS = 2.00 V
V−REF
31
V+IN
TSC7126A
V−IN
COMMON
V−
26

Fig. 87-5

PRECISION TEMPERATURE TRANSDUCER WITH REMOTE SENSOR


+15V
2
VIN
VO
6
TRIM
5
REF-02
TEMP
3
GND
4
Rs
1.5k · 5%
*
VREF
Ra
Rp
50kΩ
Rb1
Rbp
2
3
OP-02
7
4
6
Rc
VOUT
−15V

* UP TO 10 FEET OF SHIELDED 4-CONDUCTOR CABLE.

FOR THEORY OF OPERATION AND CALIBRATION PROCEDURE CONSULT APPLICATION NOTE 18, "THERMOMETER APPLICATIONS OF THE REF-02".

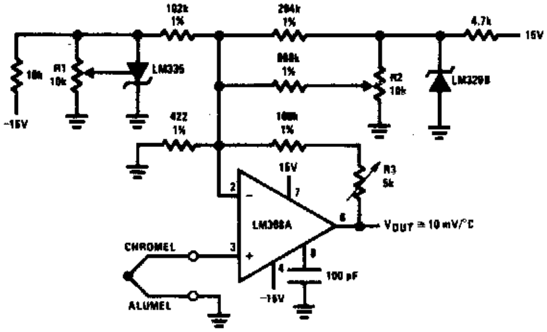
RESISTOR VALUES

TCV_{OUT} SLOPE (S)	10mV/°C	100mV/°C	10mV/°F
TEMPERATURE RANGE	−55°C to +125°C	−55°C to +125°C	−67°F to +257°C
OUTPUT VOLTAGE RANGE	−0.55V to +1.25V	−5.5V to +12.5V	−0.67V to +2.57V
ZERO SCALE	0V@0°C	0V@0°C	0V@0°F
R_a (±1% resistor)	9.09kΩ	15kΩ	7.5kΩ
R_{b1} (±% resistor)	1.5kΩ	1.82kΩ	1.21kΩ
R_{bp} (Potentiometer)	200Ω	500Ω	200Ω
R_c (±1% resistor)	5.11kΩ	84.5kΩ	8.25kΩ

* For 125°C operation, the op amp output must be able to swing to +12.5V, increase V_{IN} to +18V from +15V if this is a problem.

Fig. 87-6

CENTIGRADE CALIBRATED THERMOCOUPLE THERMOMETER


102k
1%
294k
1%
4.7k
15V
10k
R1
10k
LM335
−15V
808k
1%
R2
10k
LM329B
422
1%
100k
1%
15V
R3
5k
2
7
LM308A
6
V_{OUT} ≅ 10 mV/°C
CHROMEL
3
8
4
100 pF
−15V
ALUMEL

Terminate thermocouple reference junction in close proximity to LM335.

Adjustments:

1. Apply signal in place of thermocouple and adjust R3 for a gain of 245.7.
2. Short non-inverting input of LM308A and output of LM329B to ground.
3. Adjust R1 so that V_{OUT} = 2.982V @ 25°C.
4. Remove short across LM329B and adjust R2 so that V_{OUT} = 246 mV @ 25°C.
5. Remove short across thermocouple.

Fig. 87-7

μP CONTROLLED DIGITAL THERMOMETER

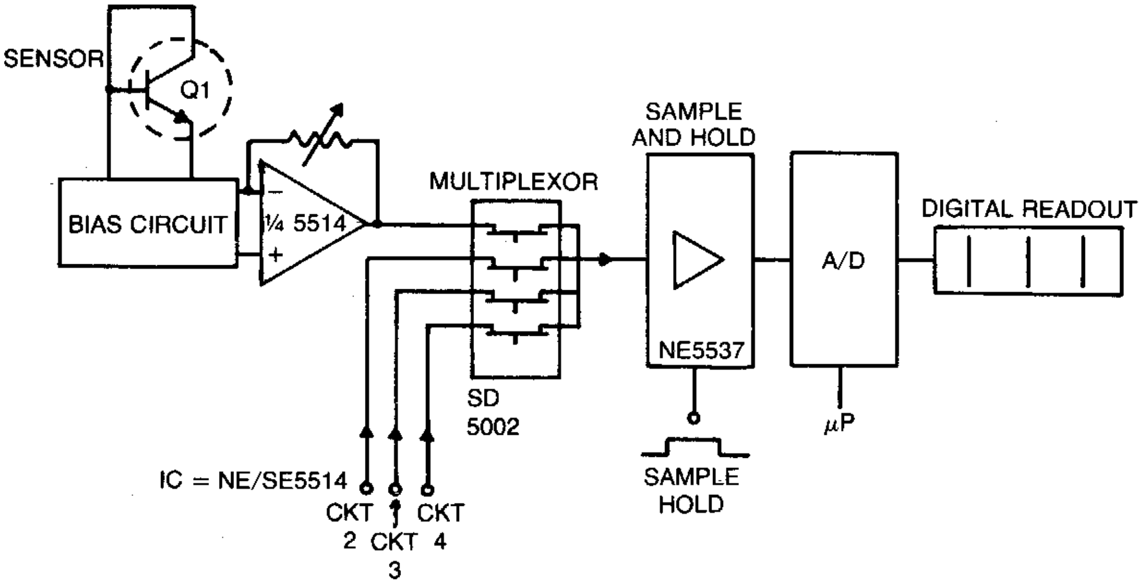

SENSOR
Q1
BIAS CIRCUIT
¼ 5514
MULTIPLEXOR
SD
5002
SAMPLE
AND HOLD
NE5537
SAMPLE
HOLD
A/D
μP
DIGITAL READOUT
IC = NE/SE5514
CKT
2
CKT
3
CKT
4

Fig. 87-8

ISOLATED TEMPERATURE SENSOR

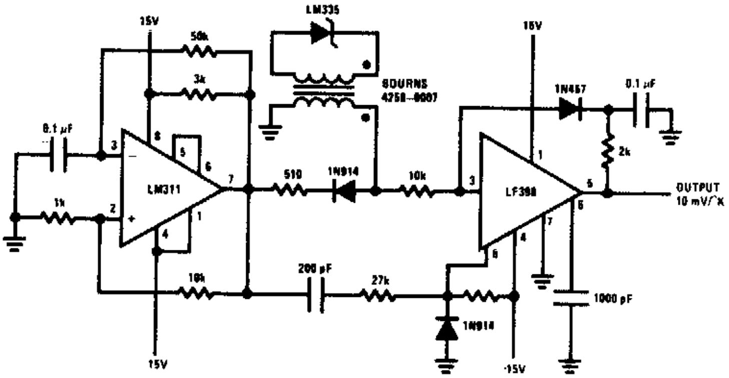

15V
50k
3k
0.1 µF
3
2
8
5
6
7
1
4
LM311
1k
10k
15V
LM335
BOURNS
4750-0007
510
1N914
10k
200 pF
27k
1N914
-15V
15V
1N457
0.1 µF
2k
LF398
3
1
5
6
7
4
8
OUTPUT
10 mV/°K
1000 pF

Fig. 87-9

DIGITAL THERMOMETER

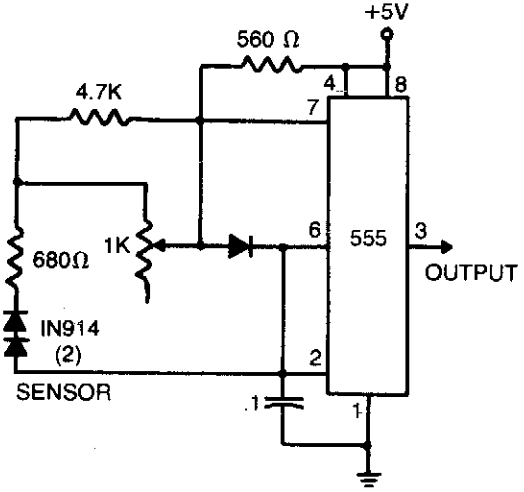
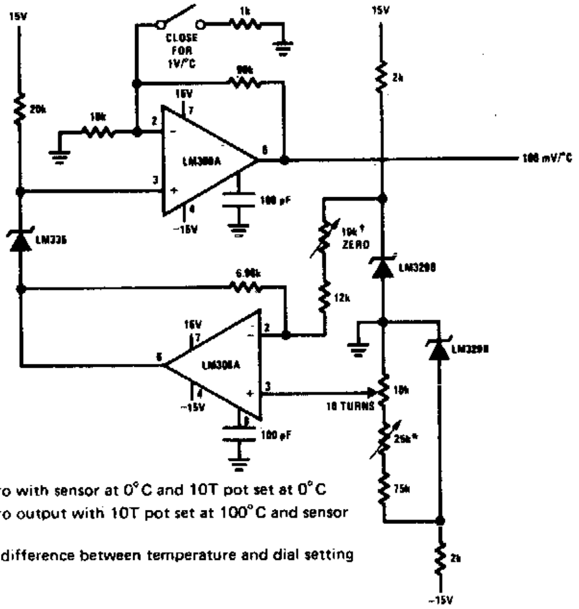

+5V
560 Ω
4.7K
4
8
7
6
555
3
OUTPUT
680Ω
1K
IN914
(2)
SENSOR
2
.1
1

Fig. 87-10

Circuit Notes

The sensor consists of two series-connected 1N914s, part of the circuit of a 555 multivibrator. Wired as shown, the output pulse rate is proportional to the temperature of the diodes. This output is fed to a simple frequency-counting circuit.

VARIABLE OFFSET THERMOMETER


15V
CLOSE FOR 1V/°C
1k
90k
20k
10k
15V
7
2
6
LM308A
3
4
−15V
100 pF
100 mV/°C
15V
2k
LM335
10k†
ZERO
LM329B
6.98k
12k
15V
7
LM308A
6
2
3
4
−15V
8
100 pF
LM329B
10k
10 TURNS
25k*
75k
2k
−15V

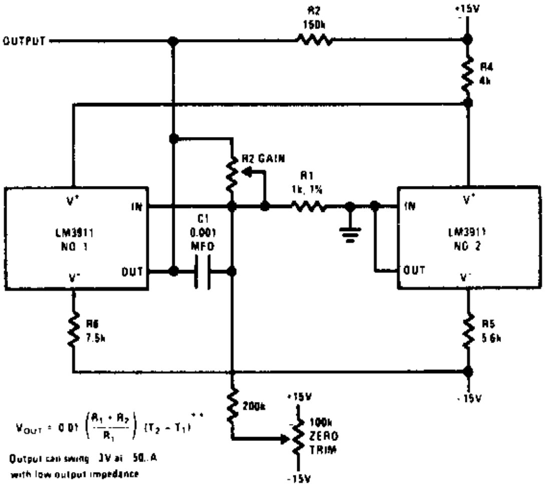
† Adjust for zero with sensor at 0°C and 10T pot set at 0°C

* Adjust for zero output with 10T pot set at 100°C and sensor at 100°C

‡ Output reads difference between temperature and dial setting of 10T pot

Fig. 87-11

DIFFERENTIAL THERMOMETER


OUTPUT
R2
150k
+15V
R4
4k
R2 GAIN
R1
1k, 1%
V+
IN
LM3911
NO 1
V−
OUT
C1
0.001
MFD
IN
V+
LM3911
NO 2
OUT
V−
R6
7.5k
R5
5.6k
−15V
200k
+15V
100k
ZERO
TRIM
−15V

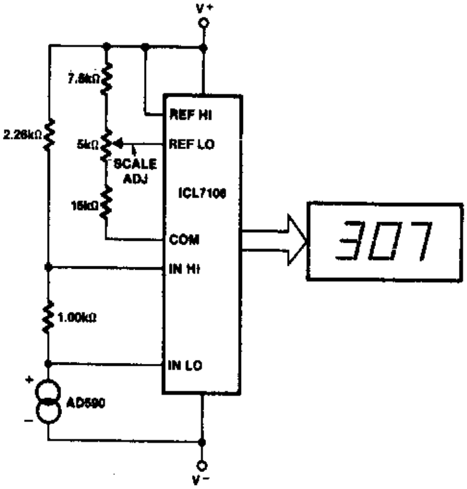
$V_{OUT} = 0.01\left(\frac{R_1 + R_2}{R_1}\right)(T_2 - T_1)^{**}$

Output can swing ±3V at ±50 μA with low output impedance

** The 0.01 in the above equation is in units of V/°K or V/°C, and is a result of the basic 0.01 V/°K sensitivity of the transducer

Fig. 87-12

BASIC DIGITAL THERMOMETER, KELVIN SCALE


V+
7.5kΩ
2.26kΩ
5kΩ
SCALE ADJ
15kΩ
REF HI
REF LO
ICL7106
COM
IN HI
1.00kΩ
IN LO
+
−
AD590
V−
307

Circuit Notes

The Kelvin scale version reads from 0 to 1999 °K theoretically, and from 223 °K to 473 °K actually. The 2.26 K resistor brings the input within the ICL7106 V_{CM} range: two general-purpose silicon diodes or an LED may be subsituted.

Fig. 87-13

BASIC DIGITAL THERMOMETER, KELVIN SCALE WITH ZERO ADJUST

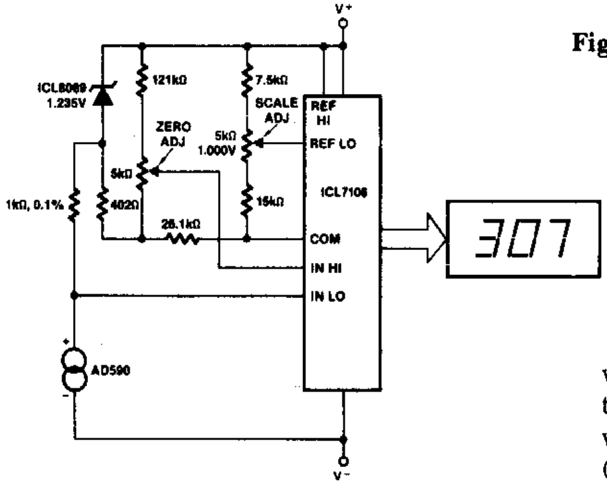
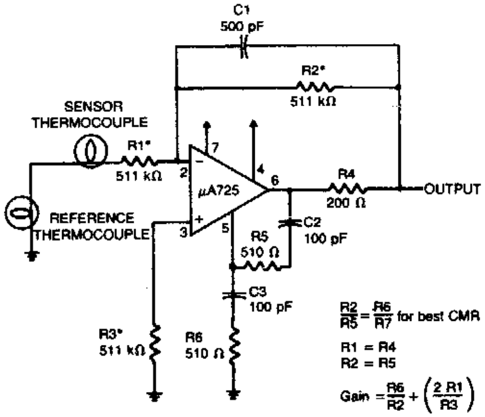

V+
ICL8069
1.235V
121kΩ
7.5kΩ
SCALE ADJ
ZERO ADJ
5kΩ
1.000V
5kΩ
402Ω
1kΩ, 0.1%
15kΩ
28.1kΩ
REF HI
REF LO
ICL7106
COM
IN HI
IN LO
+
−
AD590
V−
307

Fig. 87-14

Circuit Notes

This circuit allows zero adjustment as well as slope adjustment. The ICL8069 brings the input within the common-mode range, while the 5 K pots trim any offset at 218 °K (−55 °C), and set scale factor.

THERMOCOUPLE AMPLIFIER


C1
500 pF
R2*
511 kΩ
SENSOR
THERMOCOUPLE
R1*
511 kΩ
2
7
4
6
μA725
R4
200 Ω
OUTPUT
REFERENCE
THERMOCOUPLE
3
5
R5
510 Ω
C2
100 pF
C3
100 pF
R3*
511 kΩ
R6
510 Ω

$$\frac{R2}{R5} = \frac{R6}{R7} \text{ for best CMR}$$

R1 = R4
R2 = R5

$$\text{Gain} = \frac{R6}{R2} + \left(\frac{2\,R1}{R3}\right)$$

DC GAINS = 1000
BANDWIDTH = DC TO 540 Hz
EQUIVALENT INPUT NOISE = 0.24 μVrms

Notes
*Indicates ± 1% metal film resistors recommended for temperature stability.
Pin numbers are shown for metal package only.

Fig. 87-15

REMOTE TEMPERATURE SENSING

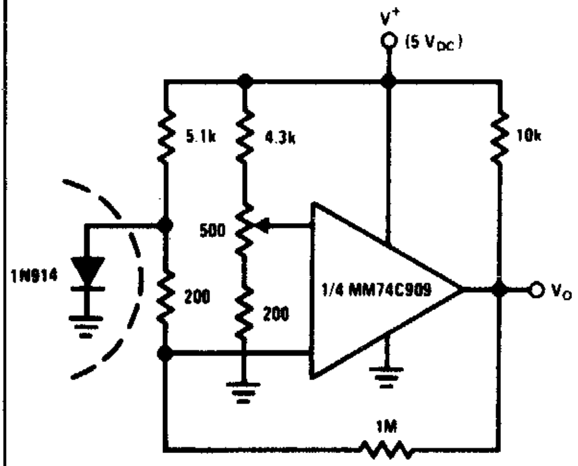
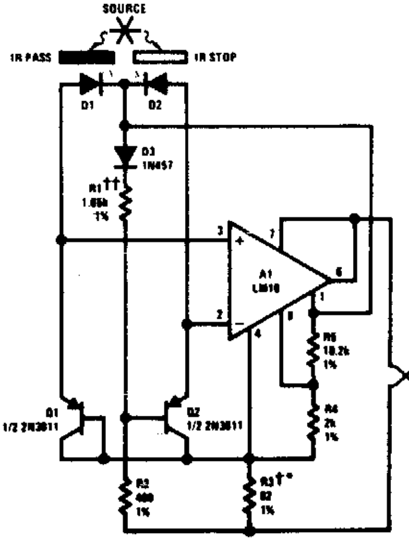

V+
(5 V$_{DC}$)
5.1k
4.3k
10k
500
1N914
200
200
1/4 MM74C909
V$_O$
1M

Fig. 87-17

OPTICAL PYROMETER


SOURCE
IR PASS
IR STOP
D1
D2
D3
1N457
R1 ††
1.05k
1%
3
7
A1
LM10
6
1
2
8
4
R5
10.2k
1%
R4
2k
1%
I$_{OUT}$
Q1
1/2 2N3811
Q2
1/2 2N3811
R2
400
1%
R3 †*
82
1%

†† Level-shift trim
* Scale factor trim
† Copper wire wound

$1\ mA \le I_{OUT} \le 5\ mA$

$0.01 \le \frac{I_{D2}}{I_{D1}} \le 100$

Fig. 87-16

SIMPLE DIFFERENTIAL TEMPERATURE SENSOR

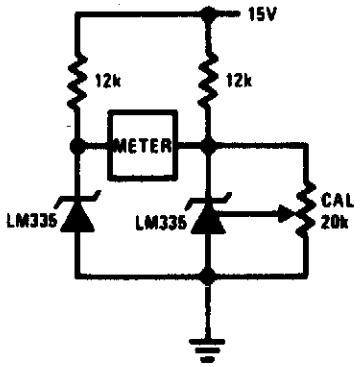

15V
12k
12k
METER
LM335
LM335
CAL
20k

Fig. 87-18

DIFFERENTIAL TEMPERATURE SENSOR

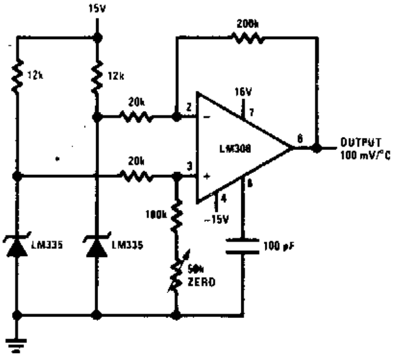
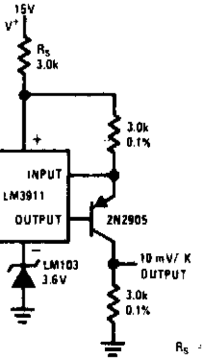

15V
200k
12k
12k
20k
20k
16V
2
7
3
LM308
6
OUTPUT
100 mV/°C
4
8
−15V
180k
50k
ZERO
100 pF
LM335
LM335

Fig. 87-19

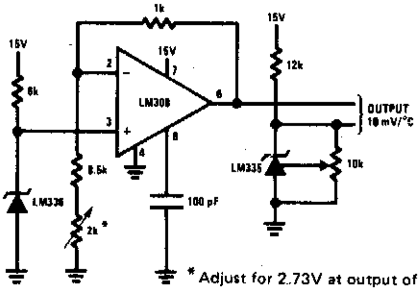
KELVIN THERMOMETER WITH GROUND REFERRED OUTPUT


15V
V+
Rs
3.0k
3.0k
0.1%
+
INPUT
LM3911
OUTPUT
2N2905
−
LM103
3.6V
10 mV/ K
OUTPUT
3.0k
0.1%

$R_S = \frac{V_S^+ - 6.8V \times 10^3 \Omega}{2}$

Fig. 87-22

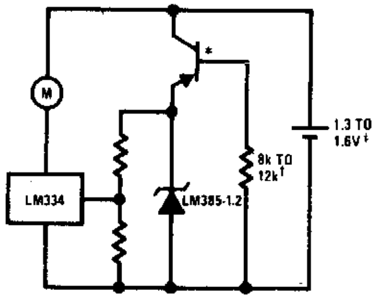
CENTIGRADE THERMOMETER


1k
15V
15V
15V
12k
8k
2
7
LM308
6
3
+
4
8
OUTPUT
10 mV/°C
8.5k
2k*
100 pF
LM335
LM335
10k

* Adjust for 2.73V at output of LM308

Fig. 87-20

LOWER POWER THERMOMETER


M
*
1.3 TO
1.6V‡
8k TO
12k†
LM334
LM385-1.2

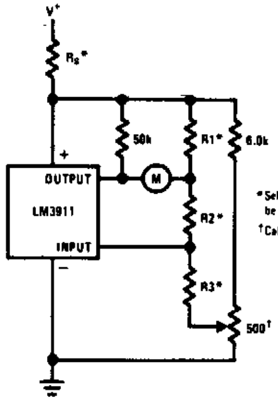
* 2N3638 or 2N2907 select for inverse $H_{FE} \cong 5$

† Select for operation at 1.3V

‡ $I_Q \approx 600\ \mu A$ to $900\ \mu A$

Fig. 87-23

METER THERMOMETER WITH TRIMMED OUTPUT


V+
Rs*
50k
R1*
6.0k
+
OUTPUT
M
LM3911
R2*
INPUT
−
R3*
500†

*Selected as for meter thermometer except T_O should be 5°K more than desired and I_O = 100μA

†Calibrates T_O

Fig. 87-21

0 °F-50 °F THERMOMETER

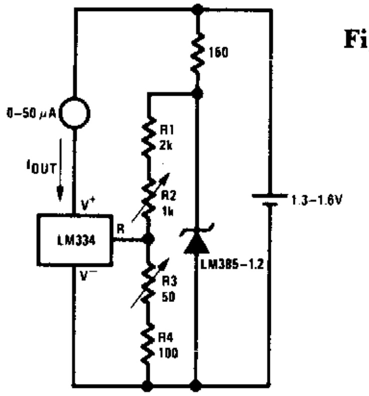

150
0–50 μA
R1
2k
IOUT
V+
R2
1k
1.3–1.6V
R
LM334
V−
LM385–1.2
R3
50
R4
100

Fig. 87-24

Calibration

1. Short LM385-1.2, adjust R3 for I_{OUT} = temp at 1.8 μA/°K
2. Remove short, adjust R2 for correct reading in °F

TEMPERATURE-TO-FREQUENCY CONVERTER

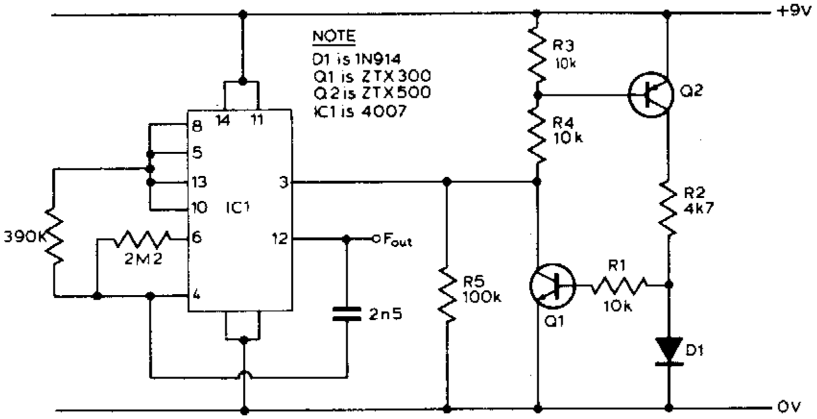
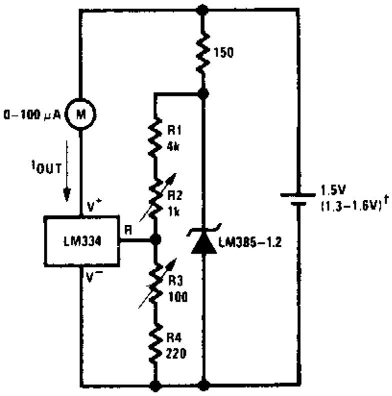

+9V
NOTE
D1 is 1N914
Q1 is ZTX300
Q2 is ZTX500
IC1 is 4007
R3
10k
R4
10k
Q2
R2
4k7
8
14
11
5
13
3
10
IC1
390K
6
12
2M2
Fout
4
R5
100k
2n5
R1
10k
Q1
D1
0V

Fig. 87-25

Circuit Notes

The circuit exploits the fact that when fed from a constant current source, the forward voltage of a silicon diode varies with temperature in a reasonably linear way. Diode D1 and resistor R2 form a potential divider fed from the constant current source. As the temperature rises, the forward voltage of D1 falls tending to turn Q1 off. The output voltage from Q1 will thus rise, and this is used as the control voltage for the CMOS VCO. With the values shown, the device gave an increase of just under 3 Hz/°C (between 0 °C and 60 °C) giving a frequency of 470 Hz at 0 °C.

0 °C-100 °C THERMOMETER


150
0–100 μA
M
I_{OUT}
R1
4k
R2
1k
1.5V
(1.3–1.6V)†
V+
LM334
R
LM385–1.2
V−
R3
100
R4
220

Calibration

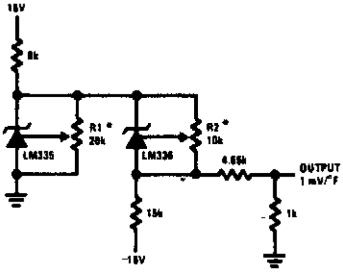
1. Short LM385-1.2, adjust R3 for I_{OUT} = temp at 1 μA/°K
2. Remove short, adjust R2 for correct reading in centigrade

† I_Q at 1.3V ≅ 500 μA

I_Q at 1.6V ≅ 2.4 mA

Fig. 87-26

GROUND REFERRED FAHRENHEIT THERMOMETER


15V
6k
R1 *
20k
LM335
R2 *
10k
LM336
4.65k
OUTPUT
1 mV/°F
15k
−15V
1k

*Adjust R2 for 2.554V across LM336.
Adjust R1 for correct output.

Fig. 87-27

GROUND REFERRED CENTIGRADE THERMOMETER

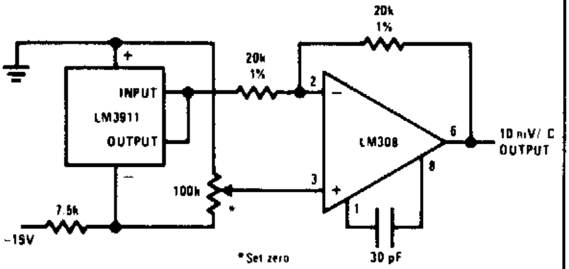

+
INPUT
LM3911
OUTPUT
−
20k
1%
20k
1%
2
−
LM308
6
10 mV/°C
OUTPUT
3
+
8
1
100k
*
7.5k
−15V
*Set zero
30 pF

Fig. 87-28

GROUND REFERRED CENTIGRADE THERMOMETER

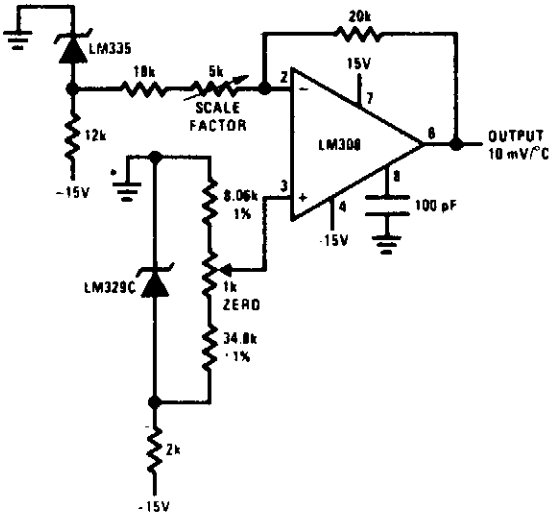

LM335
20k
18k
5k
SCALE
FACTOR
15V
7
2
−
LM308
6
OUTPUT
10 mV/°C
12k
−15V
+
8.06k
1%
3
+
8
4
100 pF
−15V
LM329C
1k
ZERO
34.8k
1%
−
2k
−15V

Fig. 87-29

TEMPERATURE SENSOR

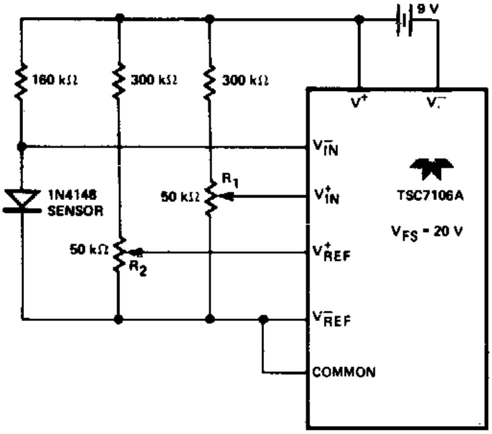

9 V
160 kΩ
300 kΩ
300 kΩ
V^{+}
V^{-}
V_{IN}^{-}
R_1
1N4148
SENSOR
50 kΩ
V_{IN}^{+}
TSC7106A
V_{FS} = 20 V
50 kΩ
R_2
V_{REF}^{+}
V_{REF}^{-}
COMMON

Fig. 87-30

POSITIVE TEMPERATURE SENSOR COEFFICIENT RESISTOR

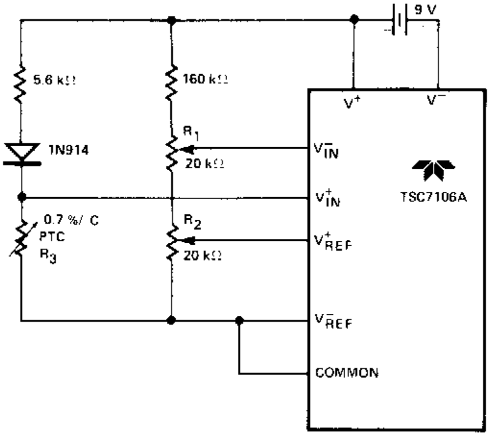
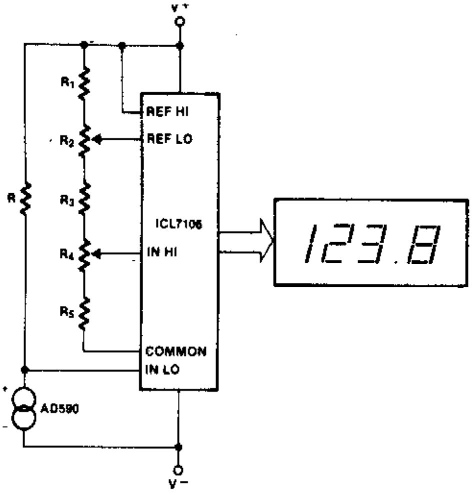

9 V
5.6 kΩ
160 kΩ
V^{+}
V^{-}
R_1
1N914
V_{IN}^{-}
20 kΩ
V_{IN}^{+}
TSC7106A
0.7 %/°C
PTC
R_3
R_2
V_{REF}^{+}
20 kΩ
V_{REF}^{-}
COMMON

Fig. 87-31

BASIC DIGITAL THERMOMETER (CELSIUS AND FAHRENHEIT SCALES)


V+
R1
R2
R3
R4
R5
R
REF HI
REF LO
ICL7106
IN HI
COMMON
IN LO
123.8
AD590
V−

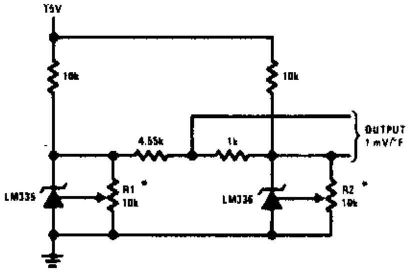
	R	R_1	R_2	R_3	R_4	R_5
°F	9.00	4.02	2.0	12.4	10.0	0
°C	5.00	4.02	2.0	5.11	5.0	11.8

Fig. 87-32

Circuit Notes

Maximum reading on the Celsius range is 199.9 °C, limited by the (short-term) maximum allowable sensor temperature. Maximum reading on the Fahrenheit range is 199.9 °F (93.3 °C), limited by the number of display digits. V_{REF} for both scales is 500 mV.

FAHRENHEIT THERMOMETER


15V
10k
10k
4.55k
1k
OUTPUT
1 mV/°F
LM335
R1 *
10k
LM336
R2 *
10k

*To calibrate adjust R2 for 2.554V across LM336.
Adjust R1 for correct output.

Fig. 87-33

88

Timers

The sources of the following circuits are contained in the Sources section beginning on page 730. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

Thumbwheel Programmable Interval Timer
Sequential Timer
Sequential Timer
Sequential UJT Timer Circuit
Time-Delayed Relay
0.1 to 90 Second Timer
Sequential Timing
Solid-State Timer for Industrial Applications
Precision Solid State Time Delay Circuit
Electronic Egg Timer
On/Off Controller
Timing Circuit
Simple Timer
Long Interval RC Timer
Timer
741 Timer
Washer Timer
Simple Time Delay

THUMBWHEEL PROGRAMMABLE INTERVAL TIMER

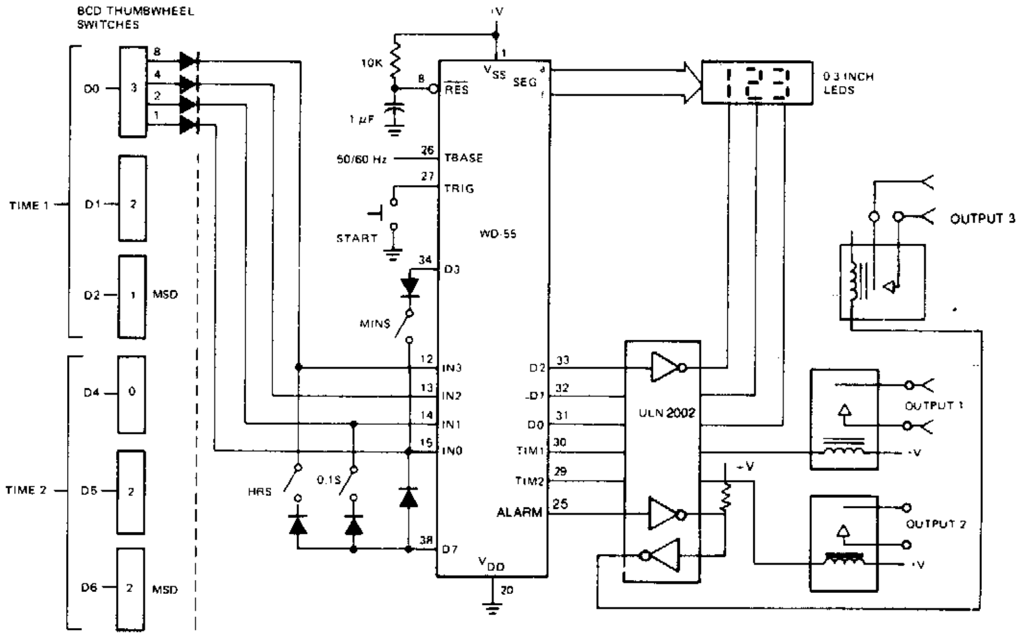
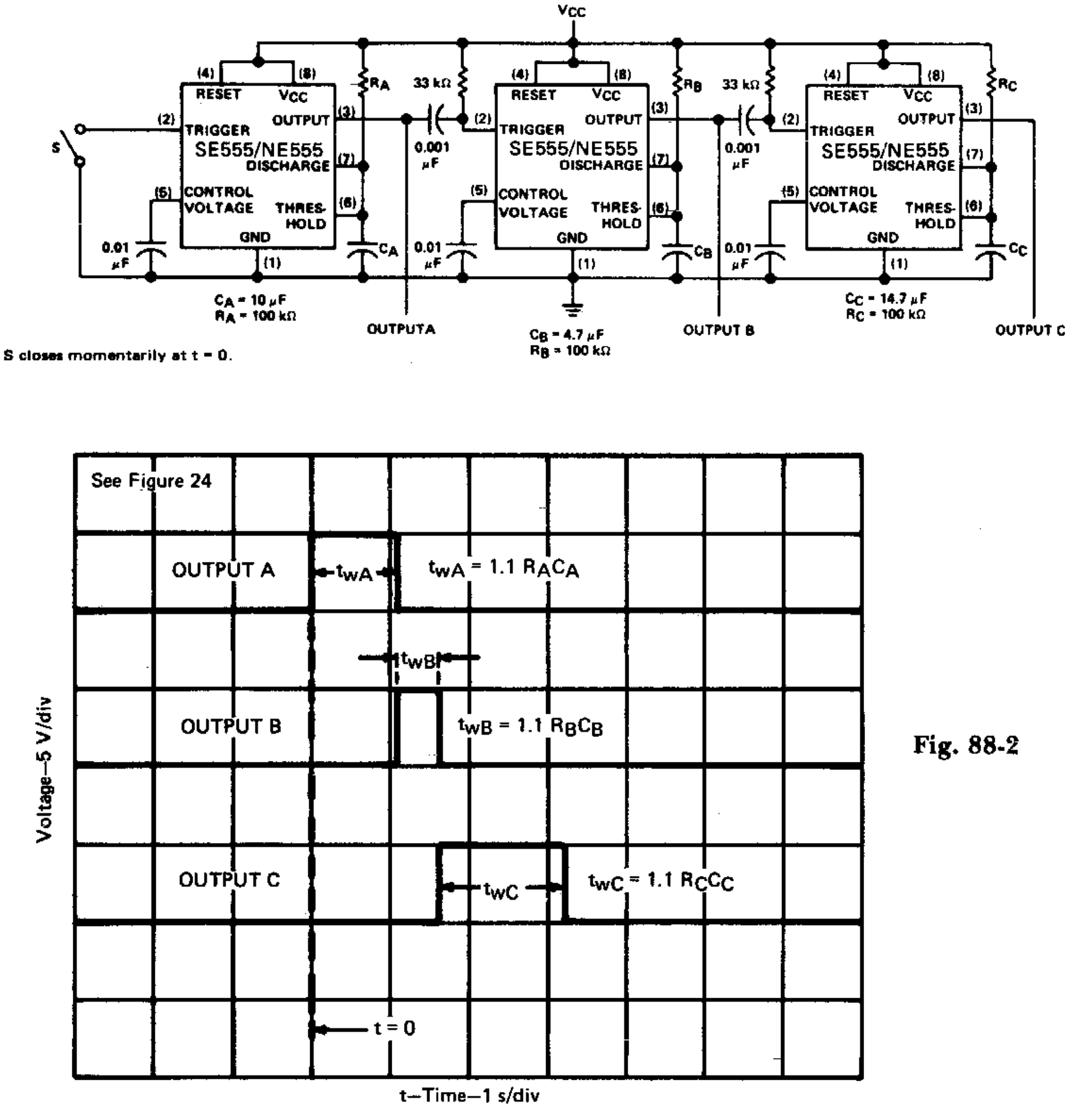

BCD THUMBWHEEL SWITCHES
+V
10K
1 μF
8
RES
1
VSS
SEG
a
f
0.3 INCH LEDS
OUTPUT 3
50/60 Hz
26
TBASE
27
TRIG
START
WD-55
34
D3
MINS
12
IN3
13
IN2
14
IN1
15
IN0
HRS
0.1S
38
D7
VDD
20
D2
33
D1
32
D0
31
TIM1
30
TIM2
29
ALARM
25
ULN 2002
OUTPUT 1
OUTPUT 2
+V
D0
D1
D2
D4
D5
D6
3
2
1
0
2
2
MSD
MSD
8
4
2
1
TIME 1
TIME 2

Fig. 88-1

Circuit Notes

Switch programmable on/off or interval timer, has three relay-switched outputs. Output one is active for the duration of time 1, output two is active for the duration of time 2, and output three is active for the duration of both one and two. Timing data is input through 6 BCD-encoded thumbwheel switches. Three SPST switches inform the WD-55 to interpret this data as NNN seconds. NNN seconds, NNN minutes, or NNN hours. The LED display will show the time remaining and the countdown when operating. Since the data is input through switches, the display may be deleted. Also, since the timing information is read from switches, the data is nonvolatile and no battery backup is required.

SEQUENTIAL TIMER



VCC
(4) RESET
(8) VCC
(2) TRIGGER
OUTPUT (3)
SE555/NE555
DISCHARGE (7)
(5) CONTROL VOLTAGE
THRESHOLD (6)
GND (1)
S
0.01 µF
RA
33 kΩ
0.001 µF
CA
RB
CB
RC
CC
CA = 10 µF
RA = 100 kΩ
CB = 4.7 µF
RB = 100 kΩ
CC = 14.7 µF
RC = 100 kΩ
OUTPUT A
OUTPUT B
OUTPUT C
See Figure 24
t_{wA}
$t_{wA} = 1.1\ R_AC_A$
t_{wB}
$t_{wB} = 1.1\ R_BC_B$
t_{wC}
$t_{wC} = 1.1\ R_CC_C$
t = 0
Voltage—5 V/div
t—Time—1 s/div

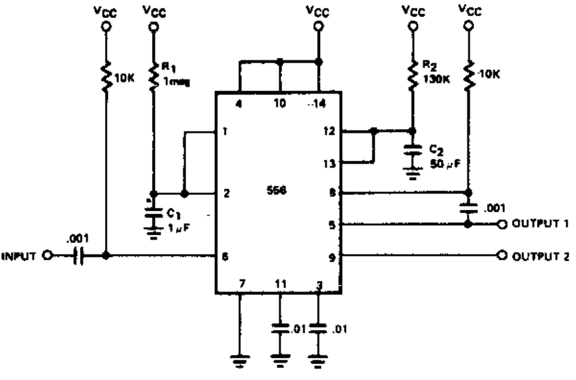
S closes momentarily at t = 0.

Fig. 88-2

Circuit Notes

Many applications, such as computers, require signals for initializing conditions during start-up. Other applications such as test equipment require activation of test signals in sequence. SE555/NE555 circuits may be connected to provide such sequential control. The timers may be used in various combinations of astable or monostable circuit connections, with or without modulation, for extremely flexible waveform control.

SEQUENTIAL TIMER


Vcc
Vcc
Vcc
Vcc
Vcc
10K
R1
1meg
R2
130K
10K
4
10
14
1
12
13
2
556
8
5
6
9
7
11
3
C2
50 µF
C1
1 µF
.001
OUTPUT 1
.001
INPUT
OUTPUT 2
.01
.01

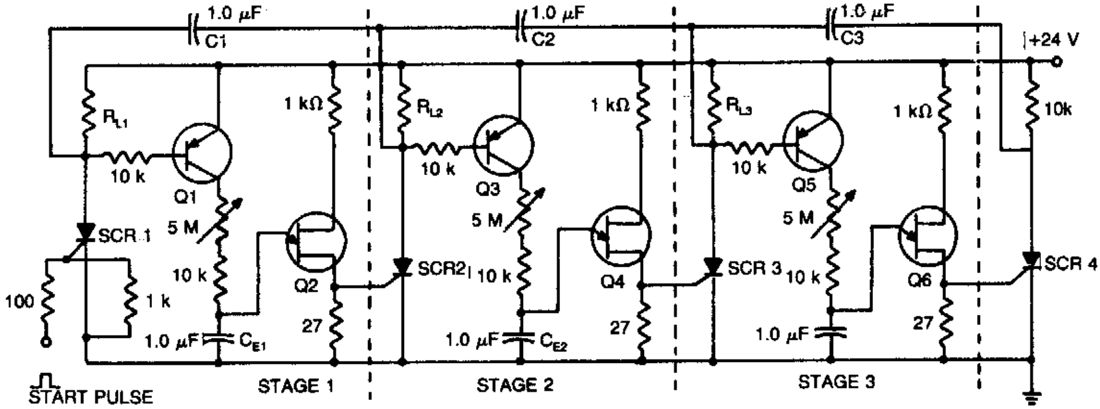
ALL RESISTOR VALUES ARE IN OHMS

Fig. 88-3

Circuit Notes

By utilizing both halves of a dual timer it is possible to obtain sequential timing. By connecting the output of the first half to the input of the second half via a .001 µF coupling capacitor sequential timing may be obtained. Delay t_1 is determined by the first half and t_2 by the second half delay. The first half of the timer is started by momentarily connecting pin 6 to ground. When it is turned out (determined by 1.1R1C1), the second half begins. Its duration is determined by 1.1R2C2.

SEQUENTIAL UJT TIMER


1.0 µF
C1
1.0 µF
C2
1.0 µF
C3
+24 V
RL1
1 kΩ
RL2
1 kΩ
RL3
1 kΩ
10k
10 k
Q1
10 k
Q3
10 k
Q5
SCR 1
5 M
5 M
5 M
Q2
Q4
Q6
SCR2
SCR 3
SCR 4
100
1 k
10 k
10 k
10 k
27
27
27
1.0 µF
CE1
1.0 µF
CE2
1.0 µF
START PULSE
STAGE 1
STAGE 2
STAGE 3
TURN OFF
CIRCUIT

UJT = 2N4853
SCR = 2N5061
TRANSISTORS = 2N4125

Fig. 88-4

TIME-DELAYED RELAY (FOR PATIO-LIGHT, GARAGE LIGHT, ENLARGER PHOTOTIMER, ETC.)

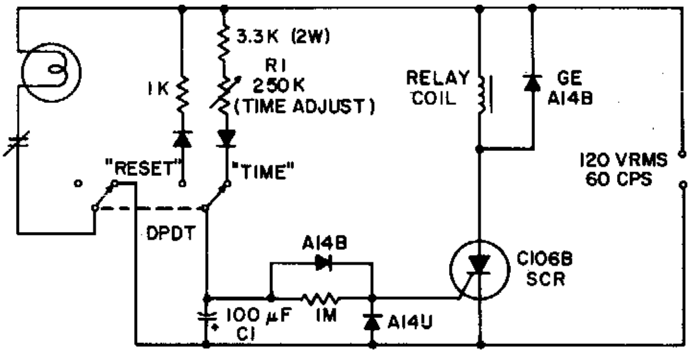

3.3K (2W)
R1
250K
(TIME ADJUST)
1K
RELAY
COIL
GE
A14B
"RESET"
"TIME"
120 VRMS
60 CPS
DPDT
A14B
C106B
SCR
100 μF
C1
1M
A14U

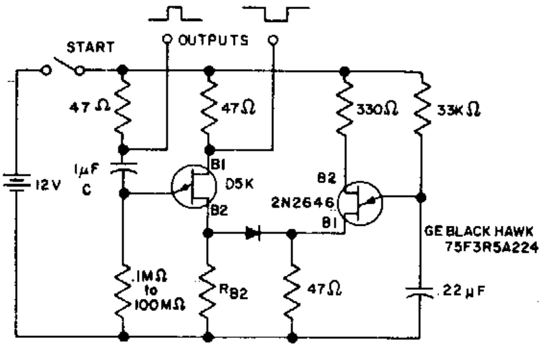
Fig. 88-5

NOTE: ALL RESISTORS 1/2 WATT

Circuit Notes

This simple timing circuit can delay an output switching function from .01 seconds to about 1 minute. The SCR is triggered by only a few microamps from the timing network R1-C1 to energize the output relay.

0.1 TO 90 SECOND TIMER


START
OUTPUTS
47Ω
47Ω
330Ω
33KΩ
1μF
C
12V
B1
D5K
B2
2N2646
B2
B1
GE BLACK HAWK
75F3R5A224
.1MΩ
to
100MΩ
R_{B2}
47Ω
.22μF

C-GE BLUE JAY AA18A105C

Fig. 88-6

Circuit Notes

The timer interval starts when power is applied to circuit and terminates when voltage is applied to load. 2N2646 is used in oscillator which pulses base 2 of D5K. This reduces the effective I_P of D5K and allows a much larger timing resistor and smaller timing capacitor to be used than would otherwise be possible.

SEQUENTIAL TIMING

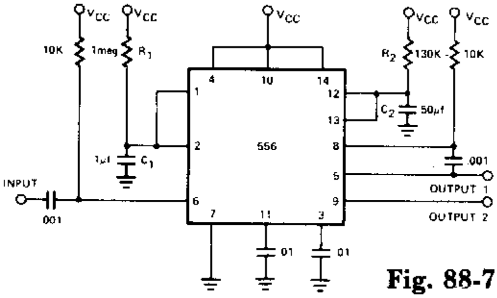

V_{CC}
V_{CC}
V_{CC}
V_{CC}
V_{CC}
10K
1meg
R_1
R_2
130K
10K
4
10
14
1
12
13
C_2
50μf
2
556
8
1μf
C_1
.001
5
INPUT
OUTPUT 1
6
9
001
7
11
3
OUTPUT 2
01
01

Fig. 88-7

Circuit Notes

By utilizing both halves of the dual timer it is possible to obtain sequential timing. By connecting the output of the first half to the input of the second half via a .001 μF coupling capacitor, sequential timing may be obtained. Delay t_1 is determined by the first half and t_2 by the second half delay. The first half of the timer is started by momentarily connecting pin 6 to ground. When it is timed out (determined by 1.1R1C1) the second half begins. Its time duration is determined by 1.1R2C2.

SOLID-STATE TIMER FOR INDUSTRIAL APPLICATIONS

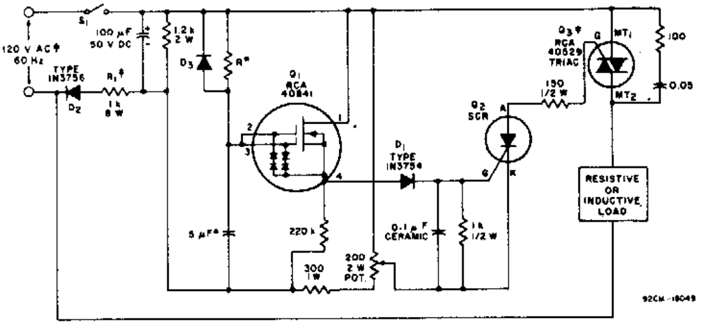

S1
120 V AC‡
60 Hz
TYPE
1N3756
D2
R1‡
1 k
8 W
100 µF
50 V DC
1.2 k
2 W
D3
R*
Q1
RCA
40841
1
2
3
4
5 µF*
220 k
300
1 W
200
2 W
POT.
D1
TYPE
1N3754
0.1 µF
CERAMIC
1 k
1/2 W
Q2
SCR
A
G
K
150
1/2 W
Q3‡
RCA
40529
TRIAC
G
MT1
MT2
100
0.05
RESISTIVE
OR
INDUCTIVE
LOAD
92CM-18049

Fig. 88-8

* Cornell-Dubilier Electronics—Type MMW or equivalent.

* R controls duration of time delay. At R = 60 MΩ up to 5-minute delay (IRC resistor, Type CGH or equivalent)

‡ This circuit can also be used at supply voltages of 240 V AC and 24V AC (60Hz) by changing the values of R1 and Q3.

TIMING CIRCUIT CHARACTERISTICS

T_A = −25°C to +60°C
Accuracy: ±10% (over temperature)
Repeatability: ±3% (at 25°C)
Reset Time: Less than 150 ms

Q2: V_{DRM} = 60V
I_{GT} = 200µA
I_T = 0.8A

D3: I_R = 1 nA
V_R = 60V

PRECISION SOLID STATE TIME DELAY CIRCUIT

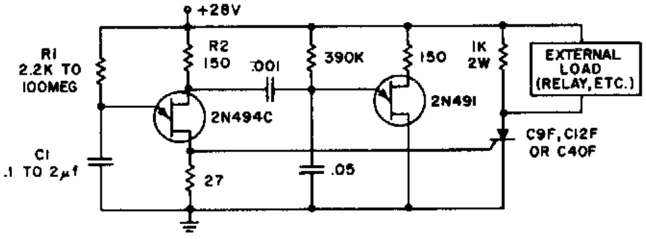
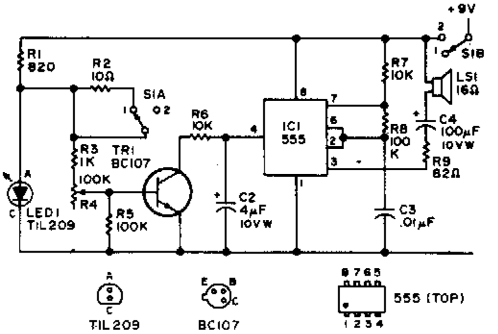

+28V
R1
2.2K TO
100MEG
C1
.1 TO 2µf
R2
150
.001
2N494C
27
390K
.05
150
2N491
1K
2W
EXTERNAL
LOAD
(RELAY, ETC.)
C9F, C12F
OR C40F

Fig. 88-9

Circuit Notes

Time delays from 0.3 milliseconds to over three minutes are possible with this circuit without using a tantalum or electrolytic capacitor. The timing interval is initiated by applying power to the circuit. At the end of the timing interval, which is determined by the value of R1C1, the 2N494C fires the controlled rectifier. This places the supply voltage minus about one volt across the load. Load currents are limited only by the rating of the controlled rectifier which is from 1 ampere up to 25 amperes for the types specified in the circuit. A calibrated potentiometer could be used in place of R1 to permit setting a predetermined time delay after one initial calibration.

ELECTRONIC EGG TIMER


+9V
2
1
S1B
R1
820
R2
10Ω
S1A
1
2
R7
10K
LS1
16Ω
8
7
6
2
3
1
4
IC1
555
R6
10K
R8
100
K
C4
100μF
10VW
R9
82Ω
R3
1K
TR1
BC107
100K
R4
A
C
LED1
TIL209
R5
100K
C2
4μF
10VW
C3
.01μF
A
C
TIL209
E
B
C
BC107
8 7 6 5
1 2 3 4
555 (TOP)

Circut Notes

The IC functions as an af multivibrator which is controlled by the external transistor. S1A/B is the on-off toggle switch.

Fig. 88-10

ON/OFF CONTROLLER

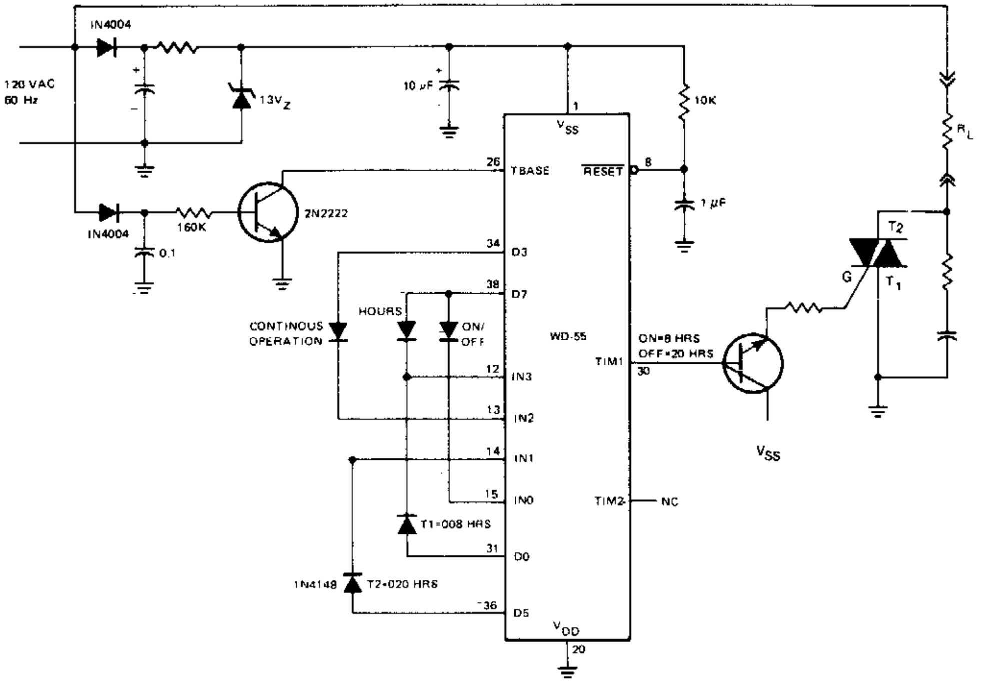
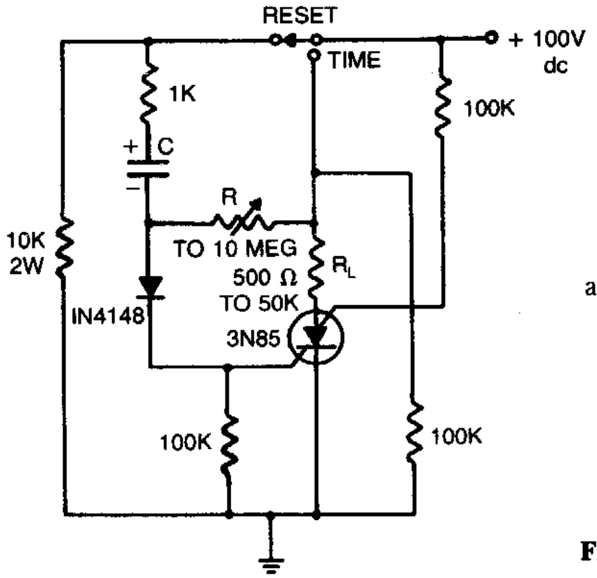

1N4004
120 VAC
60 Hz
13V_Z
10 μF
1
V_SS
10K
R_L
26
TBASE
RESET
8
1 μF
1N4004
160K
0.1
2N2222
34
D3
38
D7
HOURS
CONTINOUS
OPERATION
ON/
OFF
WD-55
TIM1
ON=8 HRS
OFF=20 HRS
30
T2
G
T1
12
IN3
13
IN2
14
IN1
15
IN0
TIM2
NC
V_SS
T1=008 HRS
31
D0
1N4148
T2=020 HRS
36
D5
V_DD
20

Fig. 88-11

Circuit Notes

The ac line-operated on/off controller is a simple, reliable solid-state alternative to a motive driven cam switch. Time 1 and time 2 are programmed by diodes to be 8 hours and 20 hours respectively. The TIM1 output is buffered by a transistor to supply gate current to a triac which switches the output load. When power is applied to the circuit, the output load is switched on for 8 hours then off for 20 hours repeatedly.

TIMING CIRCUIT


RESET
TIME
+ 100V dc
1K
+
C
−
100K
R
TO 10 MEG
10K 2W
R$_L$
500 Ω TO 50K
IN4148
3N85
100K
100K

Circuit Notes

Load current starts approximately 0.5 RC after the switch is thrown.

Fig. 88-12

SIMPLE TIMER

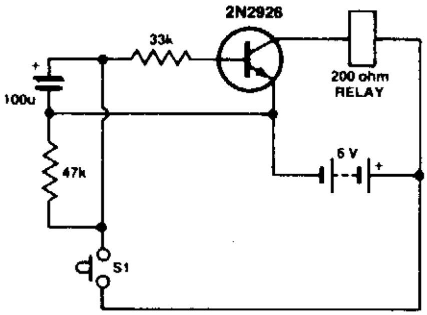
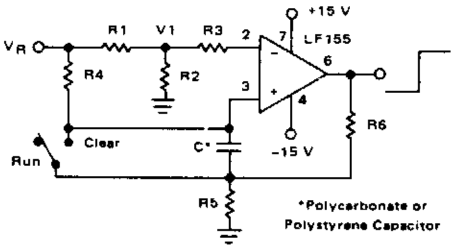

2N2926
33k
+
100u
200 ohm RELAY
6 V
+
47k
S1

Fig. 88-13

Circuit Notes

Press S1. The 100 μF electrolytic capacitor rapidly charges up at about 0.7 V. The transistor will be forward biased, and collector current will flow operating the relay. Release S1. The capacitor will begin to discharge via the 33 K resistor at the base of the transistor. When the voltage across the capacitor gets down to half a volt or so, the transistor base will no longer be forward biased, collector current will cease, and the relay will drop out. The capacitor will continue to discharge via the 47 K resistor. With the values shown, the relay will remain operated for about eight seconds. Long times are possible with lower values of capacitance by substituting a Darlington pair for the 2N2926. In this case, increase the two resitor values into the megohm range.

LONG INTERVAL RC TIMER



*Polycarbonate or Polystyrene Capacitor

Time $(t) = R4C\ell n(V_R/V_R - V_1)$, $R_3 = R_4$, $R_5 = 0.1\ R6$

If $R1 = R2$: $t = 0.693\ R4C$

Design Example: 100 Second Timer

$V_R = 10$ V C = 1 μF R3 = R4 = 144 M

R6 = 20 k R5 = 2 k R1 = R2 = 1 k

Fig. 88-14

741 TIMER

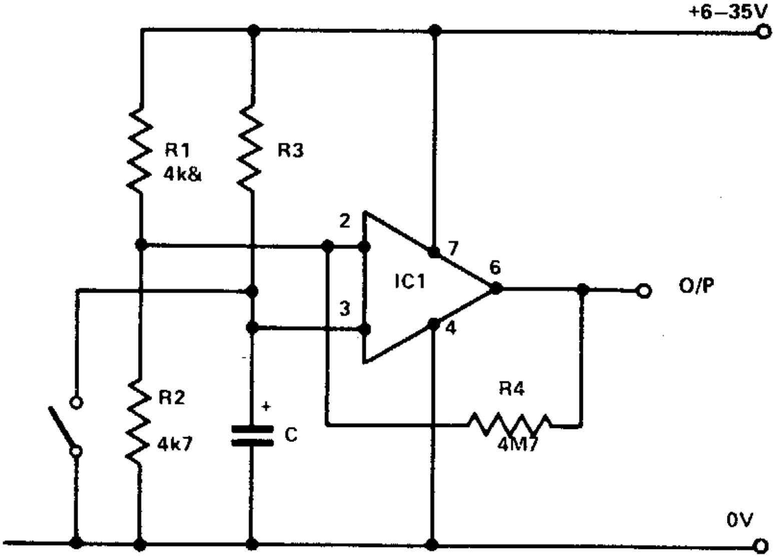


Fig. 88-15

Circuit Notes

R1 and R2 hold the inverting input at half supply voltage. R4 applies feedback to increase the input impedance at pin 3. Pin 3, the noninverting input, is connected to the junction of R3 and C. After the switch is opened, C charges via R3. When the capacitor has charged sufficiently for the potential at pin 3 to exceed that at pin 2 the output abruptly changes from 0 V to positive line potential. If reverse polarity operation is required, simply transpose R3 and C. R3 and C can be any values. Time delays from a fraction of a second to several hours can be obtained by judicious selection. The time delay—independent of supply voltage—is 0.7CR seconds where C is in farads.

TIMER

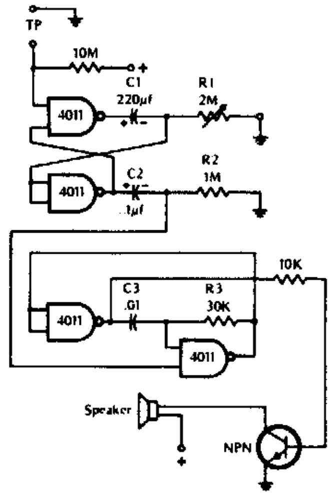
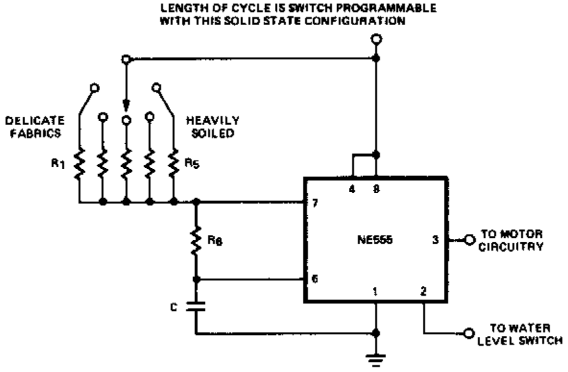

TP
10M
+
C1
220µf
R1
2M
4011
R2
1M
C2
.1µf
4011
10K
C3
.01
R3
30K
4011
4011
Speaker
+
NPN

Fig. 88-16

Circuit Notes

The timer can be used wherever time periods of up to seven minutes duration are needed. To turn on just touch the turn-on plate, and after the selected time has elapsed, an alarm will sound for a short period, then automatically turn off. The turn-on touch plate, labeled TP in the diagram, is made up of two metal strips about 1/16-inch apart. Bridging the gap with your finger activates the timer. For more time range, increase R1 and/or C1. R2 and C2 determine the period of time that the alarm will sound. Increasing either will extend the time. The tone of the alarm is determined by R3 and C3. Increasing either lowers the tone, decreasing them raises the tone.

WASHER TIMER


LENGTH OF CYCLE IS SWITCH PROGRAMMABLE WITH THIS SOLID STATE CONFIGURATION

DELICATE FABRICS
HEAVILY SOILED
R1
R5
R6
C
4
8
7
NE555
3
6
1
2
TO MOTOR CIRCUITRY
TO WATER LEVEL SWITCH

Fig. 88-17

SIMPLE TIME DELAY

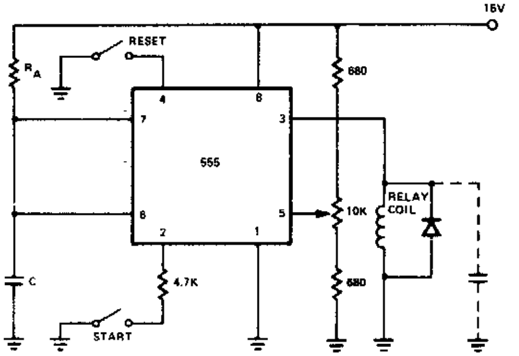

15V
RESET
RA
680
4
8
7
3
555
6
5
2
1
10K
RELAY COIL
C
4.7K
680
START

Fig. 88-18

89

Tone Controls

The sources of the following circuits are contained in the Sources section beginning on page 730. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

Stereo Phonograph Amplifier with Bass Tone Control
Equalizer
Three-Channel Tone Control
IC Preamplifier with Tone Control
Amplifier with Bass Boost
Active Bass and Treble Tone Control with Buffer
Passive Bass and Treble Tone Control
Baxendall Tone-Control Circuit
High Quality Tone Control
Microphone Preamplifier with Tone Control
Hi-Fi Tone Control Circuit
Three-Band Active Tone Control
Tone Control Circuit

STEREO PHONOGRAPH AMPLIFIER WITH BASS TONE CONTROL

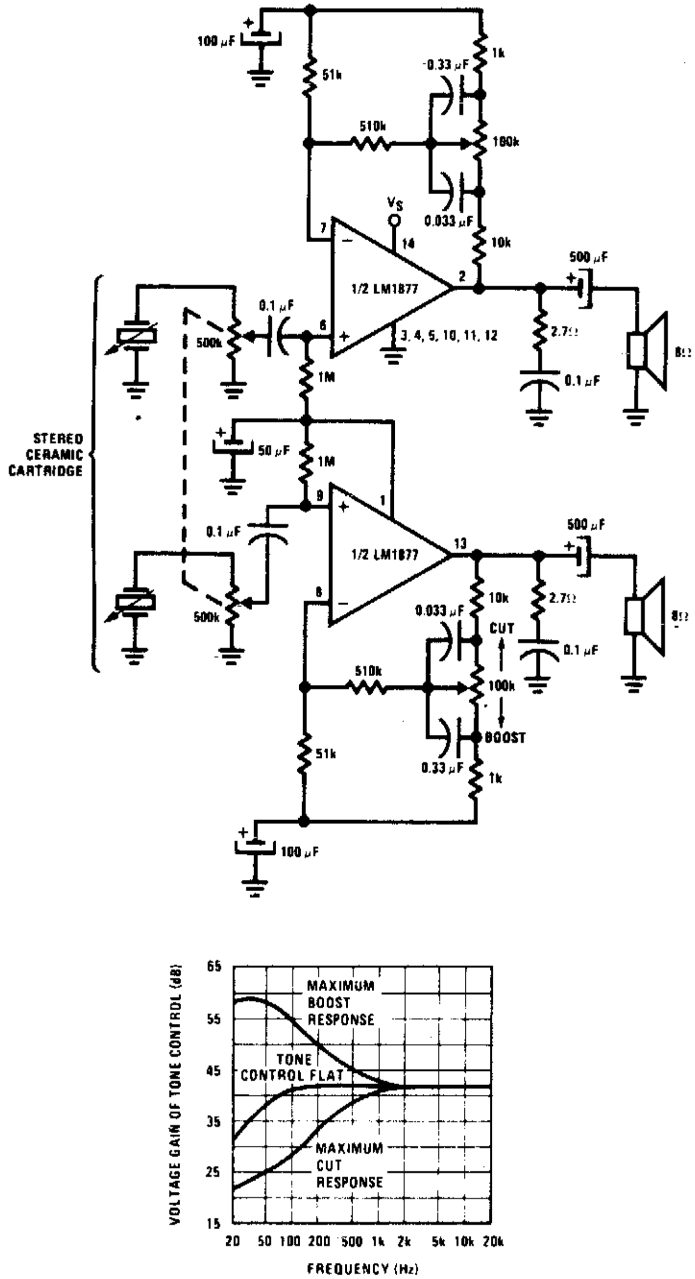
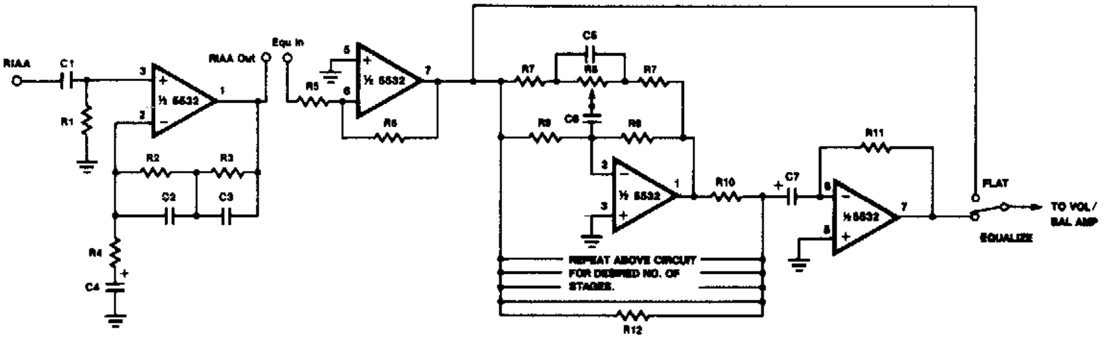

100 μF
51k
1k
0.33 μF
510k
100k
0.033 μF
Vs
14
10k
7
2
1/2 LM1877
500 μF
0.1 μF
6
3, 4, 5, 10, 11, 12
500k
2.7Ω
8Ω
1M
0.1 μF
STEREO
CERAMIC
CARTRIDGE
50 μF
1M
9
1
0.1 μF
1/2 LM1877
13
500 μF
8
500k
10k
2.7Ω
8Ω
0.033 μF
CUT
510k
100k
0.1 μF
BOOST
51k
0.33 μF
1k
100 μF
VOLTAGE GAIN OF TONE CONTROL (dB)
65
55
45
35
25
15
MAXIMUM
BOOST
RESPONSE
TONE
CONTROL FLAT
MAXIMUM
CUT
RESPONSE
20
50
100
200
500
1k
2k
5k
10k
20k
FREQUENCY (Hz)

Fig. 89-1

EQUALIZER


RIAA
C1
R1
3
2
½ 5532
1
RIAA Out
R2
R3
C2
C3
R4
C4
Equ In
R5
5
6
½ 5532
7
R6
C5
R7
R8
R7
C6
R9
R9
2
3
½ 5532
1
R10
C7
R11
6
5
½ 5532
7
FLAT
TO VOL/
BAL AMP
EQUALIZE
REPEAT ABOVE CIRCUIT
FOR DESIRED NO. OF
STAGES.
R12

COMPONENT VALUE TABLES

R8 = 25k R7 = 2.4k R9 = 240k			R8 = 50k R7 = 5.1k R9 = 510k			R8 = 100k R7 = 10k R9 = 1meg		
fo	C5	C6	fo	C5	C6	fo	C5	C6
23 Hz	1μF	.1μF	25 Hz	.47μF	.047μF	12 Hz	.47μF	.047μF
50 Hz	.47μF	.047μF	36 Hz	.33μF	.033μF	18 Hz	.33μF	.033μF
72 Hz	.33μF	.033μF	54 Hz	.22μF	.022μF	27 Hz	.22μF	.022μF
108 Hz	.22μF	.022μF	79 Hz	.15μF	.015μF	39 Hz	.15μF	.015μF
158 Hz	.15μF	.015μF	119 Hz	.1μF	.01μF	59 Hz	.1μF	.01μF
238 Hz	.1μF	.01μF	145 Hz	.082μF	.0082μF	72 Hz	.082μF	.0082μF
290 Hz	.082μF	.0082μF	175 Hz	.068μF	.0068μF	87 Hz	.068μF	.0068μF
350 Hz	.068μF	.0068μF	212 Hz	.056μF	.0056μF	106 Hz	.056μF	.0056μF
425 Hz	.056μF	.0056μF	253 Hz	.047μF	.0047μF	126 Hz	.047μF	.0047μF
506 Hz	.047μF	.0047μF	360 Hz	.033μF	.0033μF	180 Hz	.033μF	.0033μF
721 Hz	.033μF	.0033μF	541 Hz	.022μF	.0022μF	270 Hz	.022μF	.0022μF
1082 Hz	.022μF	.0022μF	794 Hz	.015μF	.0015μF	397 Hz	.015μF	.0015μF
1588 Hz	.015μF	.0015μF	1191 Hz	.01μF	.001μF	595 Hz	.01μF	.001μF
2382 Hz	.01μF	.001μF	1452 Hz	.0082μF	820pF	726 Hz	.0082μF	820pF
2904 Hz	.0082μF	820pF	1751 Hz	.0068μF	680pF	875 Hz	.0068μF	680pF
3502 Hz	.0068μF	680pF	2126 Hz	.0056μF	560pF	1063 Hz	.0056μF	560pF
4253 Hz	.0056μF	560pF	2534 Hz	.0047μF	470pF	1267 Hz	.0047μF	470pF
5068 Hz	.0047μF	470pF	3609 Hz	.0033μF	330pF	1804 Hz	.0033μF	330pF
7218 Hz	.0033μF	330pF	5413 Hz	.0022μF	220pF	2706 Hz	.0022μF	220pF
10827 Hz	.0022μF	220pF	7940 Hz	.0015μF	150pF	3970 Hz	.0015μF	150pF
15880 Hz	.0015μF	150pF	11910 Hz	.001μF	100pF	5955 Hz	.001μF	100pF
23820 Hz	.001μF	100pF	14524 Hz	820pF	82pF	7262 Hz	820pF	82pF
			17514 Hz	680pF	68pF	8757 Hz	680pF	68pF
			21267 Hz	560pF	56pF	10633 Hz	560pF	56pF
						12670 Hz	470pF	47pF
						18045 Hz	330pF	33pF

COMPONENT VALUES

R1	1meg	C1	.22μF
R2	100k	C2	750pF
R3	1meg	C3	.0033μF
R4	1.1k	C4	33μF
R5	100k	C5	SEE TABLE
R6	100k	C6	SEE TABLE
R7	SEE TABLE	C7	2.2μF
R8	(pot) SEE TABLE		
R9	SEE TABLE		
R10	100k		
R11	100k		
R12	20k (5 STAGES)		

Fig. 89-2

THREE-CHANNEL TONE CONTROL

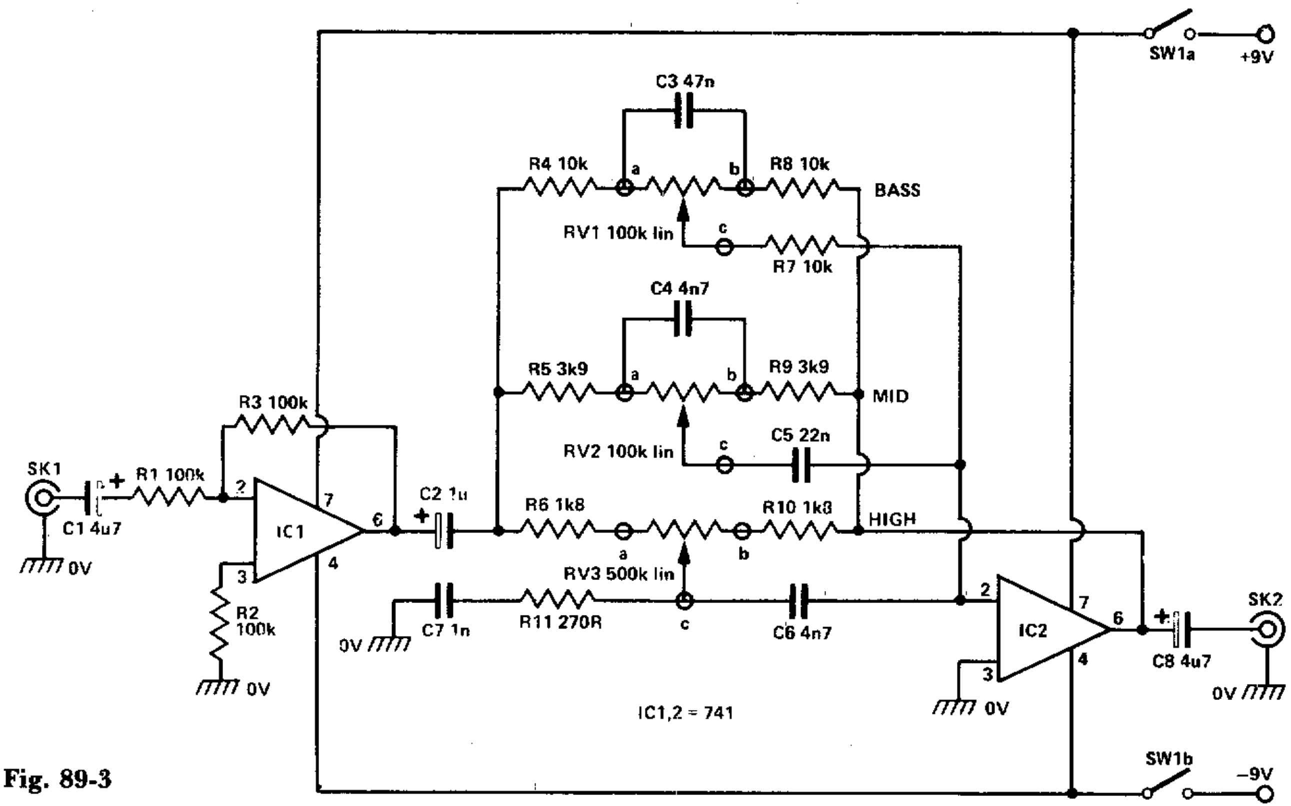

SW1a
+9V
C3 47n
R4 10k
a
b
R8 10k
BASS
RV1 100k lin
c
R7 10k
C4 4n7
R5 3k9
R9 3k9
MID
R3 100k
RV2 100k lin
C5 22n
SK1
R1 100k
2
7
C2 1u
6
R6 1k8
R10 1k8
HIGH
C1 4u7
IC1
0V
4
3
RV3 500k lin
R2
100k
0V
C7 1n
R11 270R
C6 4n7
IC2
C8 4u7
SK2
IC1,2 = 741
SW1b
-9V

Fig. 89-3

Circuit Notes

The input signal is fed via SK1 to the first active stage built around IC1. Configured as a noninverting amplifier whose gain is set by the ratio of R3 and R1. In this case, the gain is set at unity. This initial stage is required to isolate the following stage from any loading effects. The output from IC1 is fed via three frequency shaping networks to IC2. The three networks built around RV1, RV2, and RV3 are also included in the feedback path of IC2, another inverting op amp stage. The components associated with the three variable resistors are chosen to give the required frequency control.

IC PREAMPLIFIER WITH TONE CONTROL

IC PREAMPLIFIER RESPONSE CHARACTERISTICS

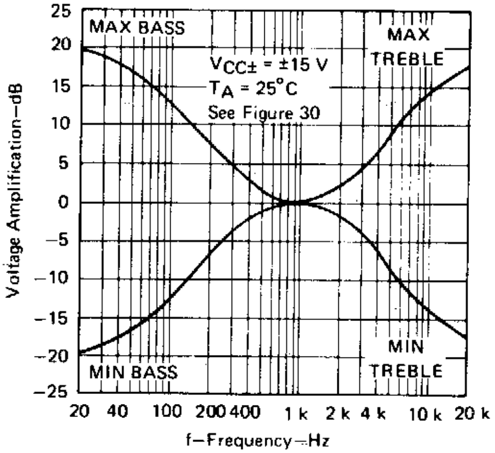
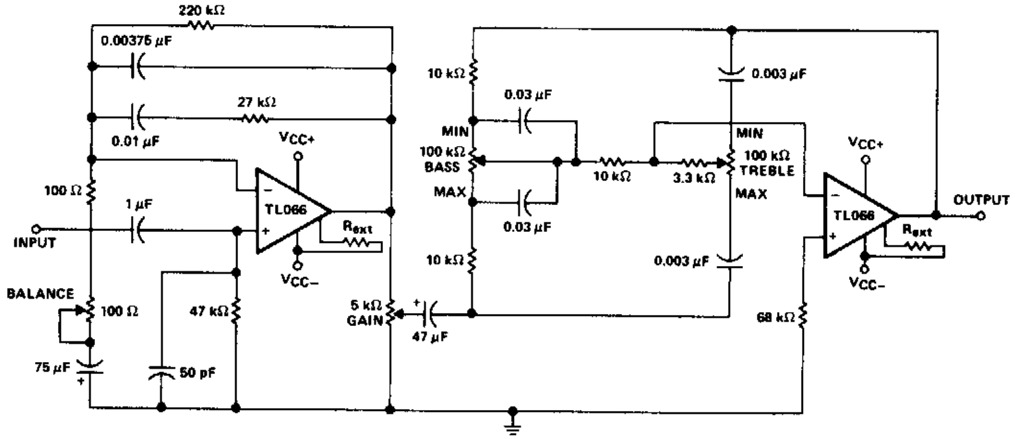
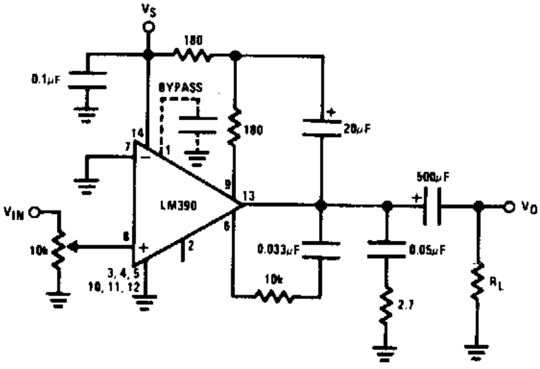


Fig. 89-4



220 kΩ
0.00375 μF
27 kΩ
0.01 μF
VCC+
100 Ω
1 μF
INPUT
TL066
Rext
VCC−
BALANCE
100 Ω
47 kΩ
75 μF
50 pF
5 kΩ
GAIN
47 μF
10 kΩ
0.03 μF
MIN
100 kΩ
BASS
MAX
10 kΩ
0.03 μF
10 kΩ
3.3 kΩ
0.003 μF
MIN
100 kΩ
TREBLE
MAX
0.003 μF
68 kΩ
VCC+
TL066
Rext
VCC−
OUTPUT

AMPLIFIER WITH BASS BOOST



VS
180
0.1μF
BYPASS
180
20μF
14
7
1
9
13
LM390
6
8
2
3, 4, 5
10, 11, 12
VIN
10k
500μF
VO
0.033μF
0.05μF
10k
2.7
RL

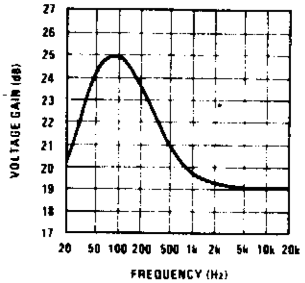


Fig. 89-5

Frequency Response with Bass Boost

ACTIVE BASS & TREBLE TONE CONTROL WITH BUFFER

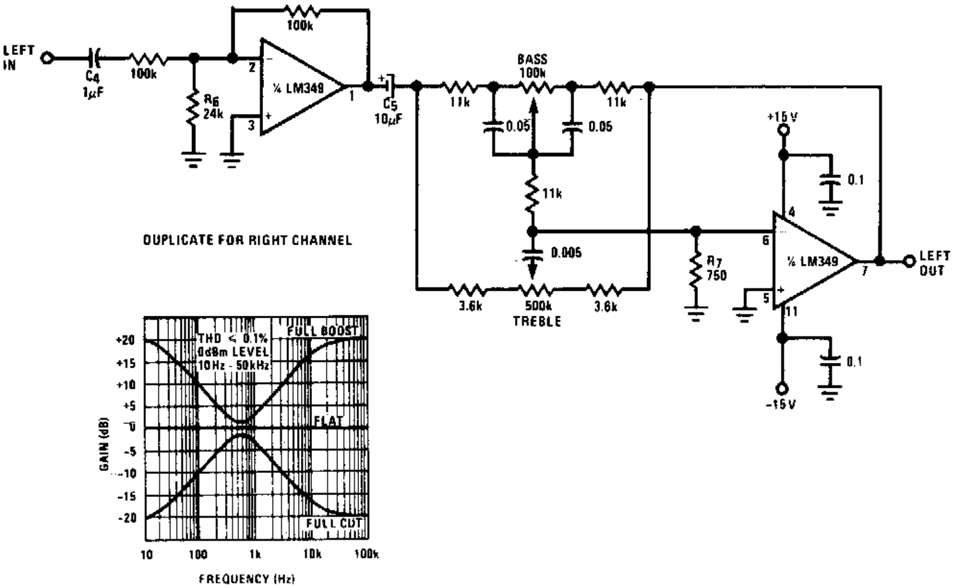
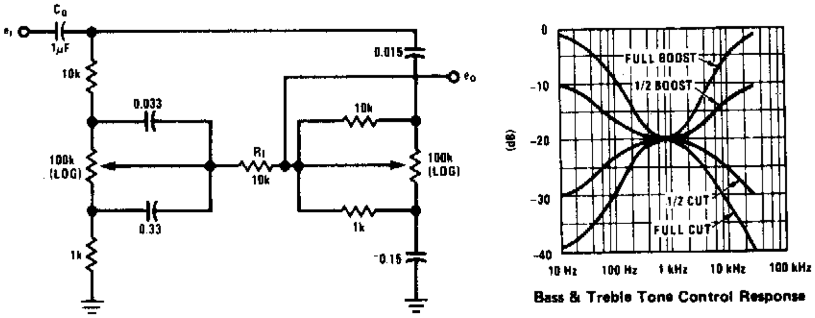

LEFT IN
C4 1μF
100k
R6 24k
100k
¼ LM349
2
3
1
C5 10μF
11k
BASS 100k
11k
0.05
0.05
11k
0.005
3.6k
500k TREBLE
3.6k
R7 750
+15 V
0.1
4
6
5
11
7
¼ LM349
LEFT OUT
0.1
-15 V
DUPLICATE FOR RIGHT CHANNEL
THD ≤ 0.1%
0dBm LEVEL
10Hz - 50kHz
FULL BOOST
FLAT
FULL CUT
GAIN (dB)
+20
+15
+10
+5
0
-5
-10
-15
-20
10
100
1k
10k
100k
FREQUENCY (Hz)

Fig. 89-6

PASSIVE BASS & TREBLE TONE CONTROL


ei
C0
1μF
10k
0.033
100k (LOG)
0.33
1k
R1
10k
0.015
eo
10k
100k (LOG)
1k
0.15
0
-10
-20
-30
-40
(dB)
FULL BOOST
1/2 BOOST
1/2 CUT
FULL CUT
10 Hz
100 Hz
1 kHz
10 kHz
100 kHz

Bass & Treble Tone Control Response

Fig. 89-7

BAXENDALL TONE-CONTROL CIRCUIT

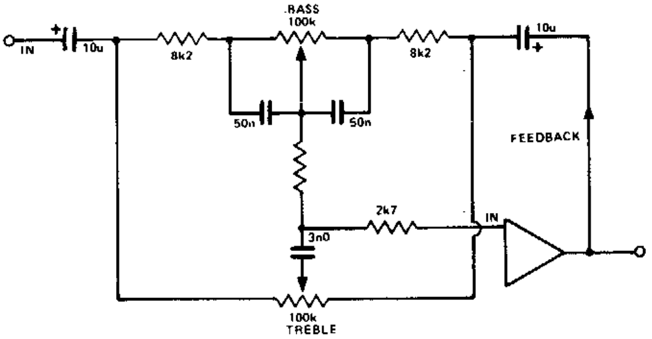

IN
10u
8k2
BASS 100k
8k2
10u
50n
50n
FEEDBACK
2k7
3n0
IN
100k TREBLE

Fig. 89-8

HIGH QUALITY TONE CONTROL

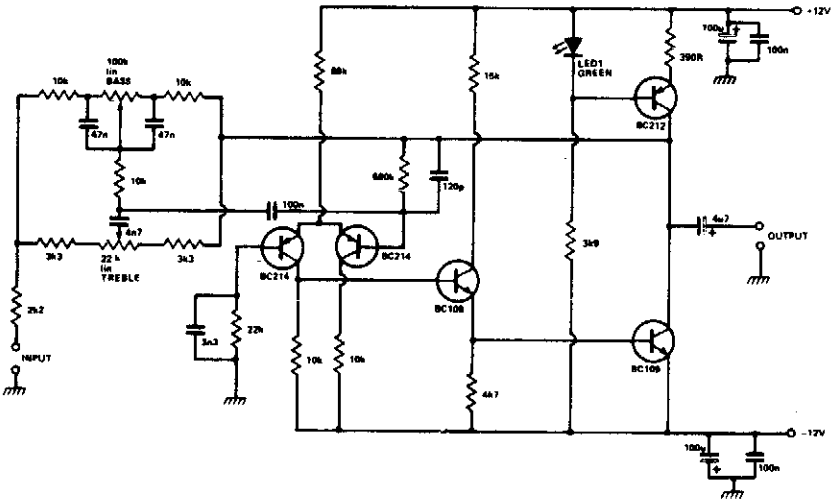

+12V
100µ
100n
390R
LED1 GREEN
BC212
88k
15k
100k lin BASS
10k
10k
47n
47n
10k
680k
120p
100n
4n7
3k3
22 k lin TREBLE
3k3
BC214
BC214
BC108
4n7
OUTPUT
3k9
2k2
INPUT
3n3
22k
10k
10k
4n7
BC109
-12V
100µ
100n

Fig. 89-9

Circuit Notes

The circuit is based on an inverting op amp using discrete transistors to overcome poor slew rate, fairly high distortion, and high noise problems. The output stage is driven by a constant current source, biased by a green LED to provide temperature compensation. With the controls flat, the unit provides unity gain so the stage can be switched in or out. The design is suitable for inputs between 100 mV and 1 V and provides a good overload margin at low distortion for the accurate reproduction of transients. The usual screening precautions against hum should be carried out.

MICROPHONE PREAMPLIFIER WITH TONE CONTROL

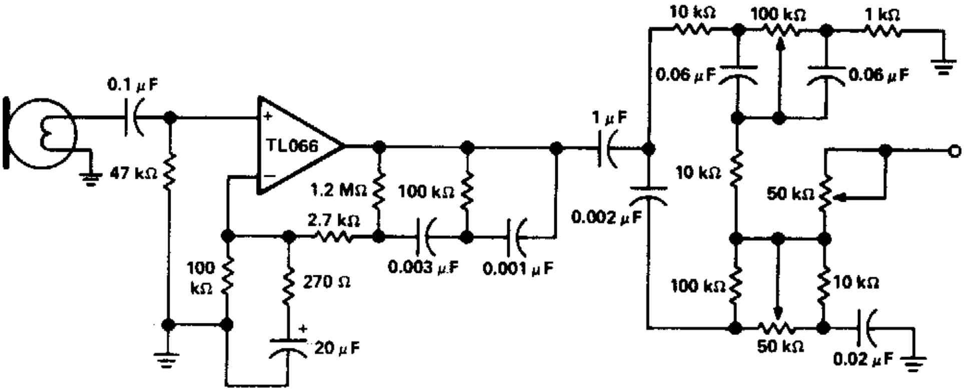

0.1 µF
47 kΩ
TL066
1.2 MΩ
100 kΩ
2.7 kΩ
100 kΩ
270 Ω
20 µF
0.003 µF
0.001 µF
1 µF
0.002 µF
10 kΩ
100 kΩ
1 kΩ
0.06 µF
0.06 µF
10 kΩ
50 kΩ
100 kΩ
10 kΩ
50 kΩ
0.02 µF

Fig. 89-10

HI-FI TONE CONTROL CIRCUIT (HIGH Z INPUT)

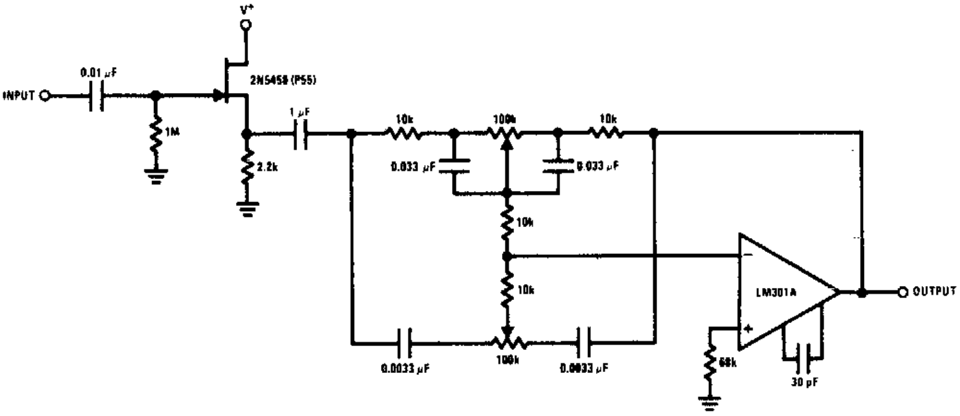

V+
INPUT
0.01 μF
1M
2N5458 (P55)
2.2k
1 μF
10k
100k
10k
0.033 μF
0.033 μF
10k
10k
LM301A
OUTPUT
0.0033 μF
100k
0.0033 μF
68k
30 pF

Fig. 89-11

Circuit Notes

The 2N5458 JFET provides the function of a high input impedance and low noise characteristics to buffer an op amp feedback tone control circuit.

THREE-BAND ACTIVE TONE CONTROL

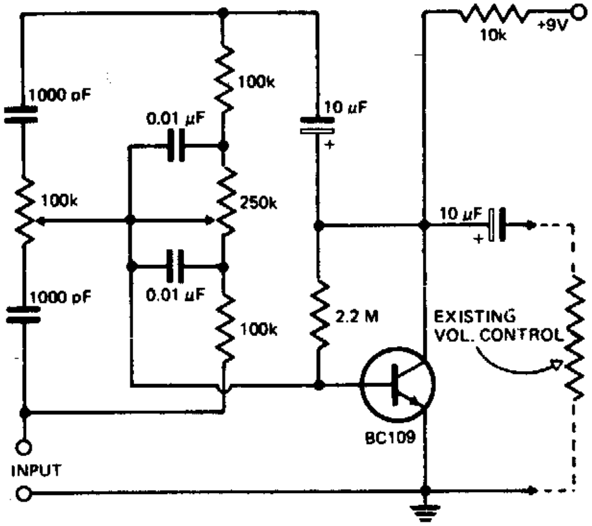
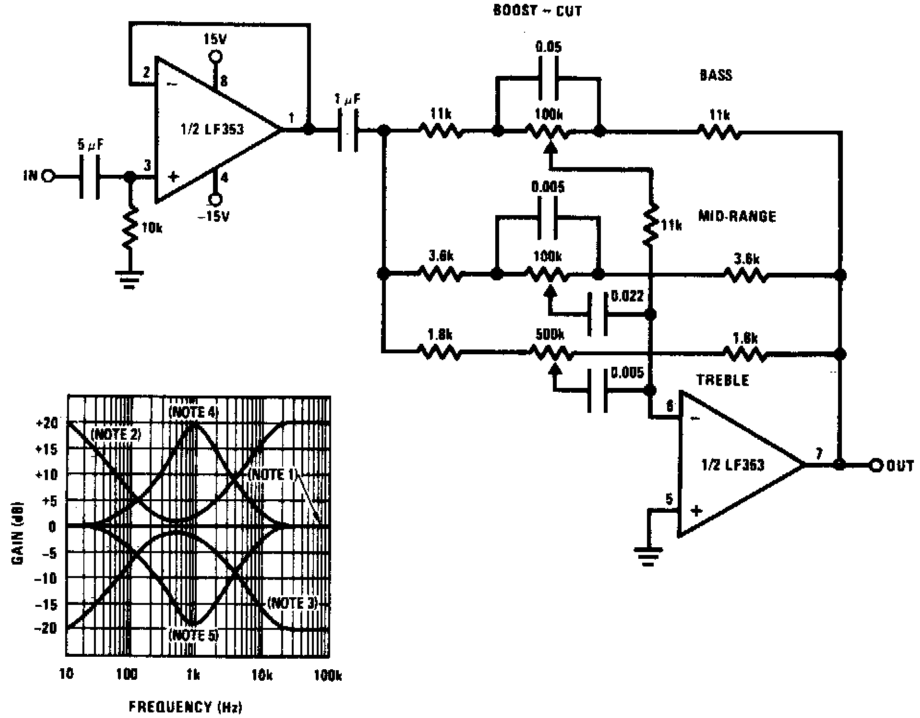

10k
+9V
1000 pF
100k
0.01 μF
100k
10 μF
+
250k
10 μF
+
0.01 μF
1000 pF
100k
2.2 M
EXISTING
VOL. CONTROL
BC109
INPUT

Fig. 89-12

TONE CONTROL CIRCUIT


BOOST ~ CUT
15V
8
2
−
1/2 LF353
1
1 μF
5 μF
IN
3
+
4
−15V
10k
0.05
11k
100k
BASS
11k
0.005
11k
MID-RANGE
3.6k
100k
3.6k
0.022
1.8k
500k
1.8k
0.005
TREBLE
6
7
1/2 LF353
OUT
5
+
+20
+15
+10
+5
0
−5
−10
−15
−20
GAIN (dB)
(NOTE 4)
(NOTE 2)
(NOTE 1)
(NOTE 3)
(NOTE 5)
10
100
1k
10k
100k
FREQUENCY (Hz)

Note 1: All controls flat.
Note 2: Bass and treble boost, mid flat.
Note 3: Bass and treble cut, mid flat.
Note 4: Mid boost, bass and treble flat.
Note 5: Mid cut, bass and treble flat.

- All potentiometers are linear taper
- Use the LF347 Quad for stereo applications

Fig. 89-13

Circuit Notes

A simple single-transistor circuit will give approximately 15 dB boost or cut at 100 Hz and 15 kHz respectively. A low noise audio type transistor is used, and the output can be fed directly into any existing amplifier volume control to which the tone control is to be fitted. The gain of the circuit is near unity when controls are set in the flat position.

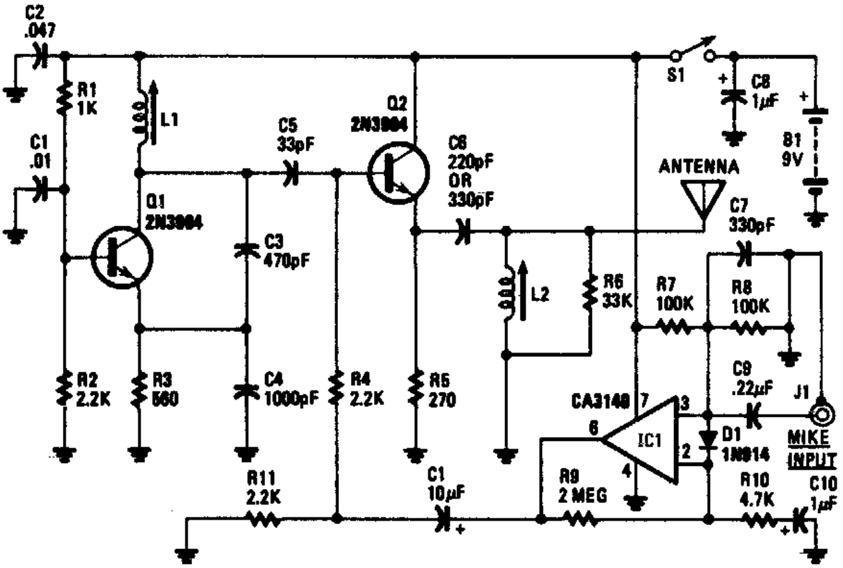
90

Transmitters

The sources of the following circuits are contained in the Sources section beginning on page 730. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

Wireless AM Microphone
27 MHz and 49 MHz RF Oscillator/Transmitter
1-2 MHz Broadcaster Transmitter
One Tube, 10 Watt C.W. Transmitter
Simple FM Transmitter

WIRELESS AM MICROPHONE



C2
.047
R1
1K
L1
C1
.01
Q1
2N3904
C5
33pF
Q2
2N3904
C6
220pF
OR
330pF
C3
470pF
L2
R6
33K
R7
100K
R8
100K
S1
C8
1μF
B1
9V
ANTENNA
C7
330pF
R2
2.2K
R3
560
C4
1000pF
R4
2.2K
R5
270
CA3140
7
3
6
IC1
2
4
C9
.22μF
J1
D1
1N914
MIKE
INPUT
R11
2.2K
C1
10μF
R9
2 MEG
R10
4.7K
C10
1μF

Fig. 90-1

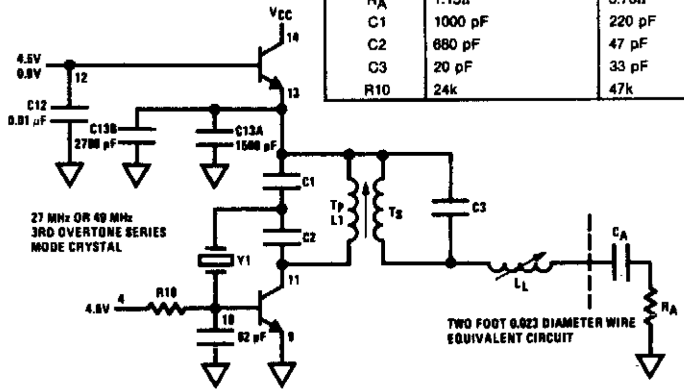
Circuit Notes

Transistor Q1 and its associated components comprise a tuneable rf oscillator. The rf signal is fed to transistor Q2, the modulator. Operational amplifier IC1 increases the audio signal and applies it through resistor R4 to the base of Q2. Tune an AM radio to an unused frequency between 800 to 1600 kHz. Tune L1 for a change in the audio level coming from the radio. Peak the output by adjusting L2. If L1 is disturbed, it may be necessary to readjust L2 for peak performance. Depending on the impedance of the microphone audio sensitivity can be increased by decreasing the value of R10 and vice versa.

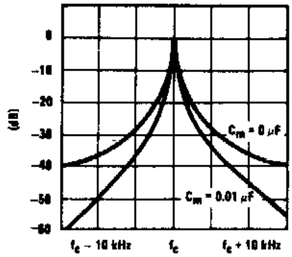
27 MHz AND 49 MHz RF OSCILLATOR/TRANSMITTER

Component	27 MHz	49 MHz
T_P	2 Turns	6 Turns
T_S	3 Turns	1 Turn
L1	TOKO KXN K4636 BJF	TOKO KEN K4635 BJE
L_L	MILLER #4611	MILLER #9336-10
C_A	5.4 pF	6.2 pF
R_A	1.15Ω	3.78Ω
C1	1000 pF	220 pF
C2	680 pF	47 pF
C3	20 pF	33 pF
R10	24k	47k

Use TOKO form #51-0116-02 and #30 wir
or #51-0178 and #32 wire



VCC
14
4.5V
0.9V
12
13
C12
0.01 μF
C13B
2700 pF
C13A
1500 pF
C1
C2
C3
Tp
L1
Ts
27 MHz OR 49 MHz
3RD OVERTONE SERIES
MODE CRYSTAL
Y1
11
4.8V
4
R10
10
82 pF
9
LL
CA
RA
TWO FOOT 0.023 DIAMETER WIRE
EQUIVALENT CIRCUIT



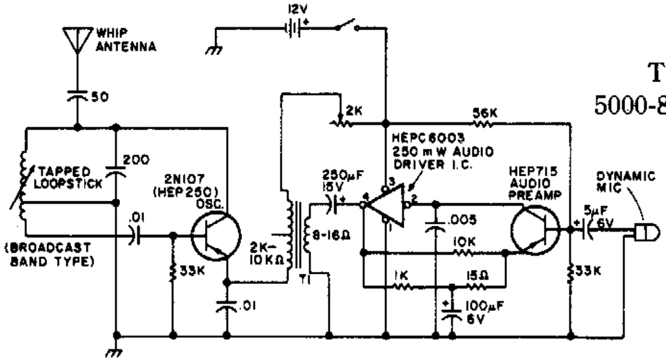
Envelope of Transmitted Spectrum for Circuit

Fig. 90-2

Circuit Notes

The modulator and oscillator consist of two NPN transistors. The base of the modulator transistor is driven by a bidirectional current source with the voltage range for the high condition limited by a saturating PNP collector to the pin 4 VREG voltage and low condition limited by a saturating NPN collector in series with a diode to ground. The crystal oscillator/transmitter transistor is configured to oscillate in a class C mode. Because third overtone crystals are used for 27 MHz or 49 MHz applications a tuned collector load must be used to guarantee operation at the correct frequency.

1-2 MHz BROADCAST TRANSMITTER



WHIP
ANTENNA
12V
50
200
TAPPED
LOOPSTICK
(BROADCAST
BAND TYPE)
.01
2N107
(HEP250)
OSC.
33K
.01
2K
2K–
10KΩ
T1
8-16Ω
250μF
15V
HEPC6003
250 m W AUDIO
DRIVER I.C.
3
4
2
1
56K
.005
10K
1K
15Ω
100μF
6V
HEP715
AUDIO
PREAMP
5μF
6V
DYNAMIC
MIC
33K

Circuit Notes

T1 is a low impedance output transformer 5000-8 ohms.

Fig. 90-3

ONE TUBE, 10 WATT C.W. TRANSMITTER

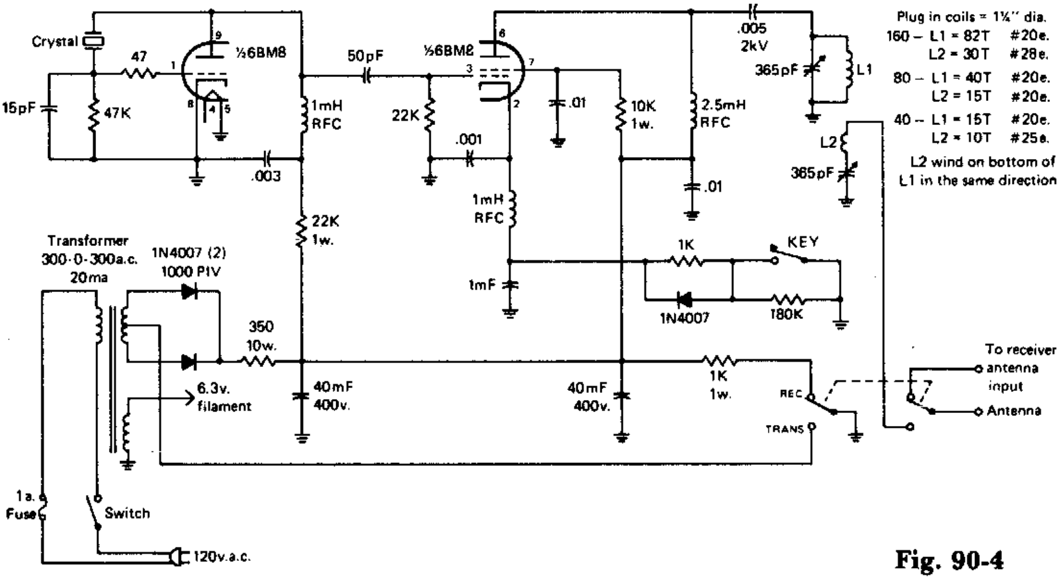

Crystal
47
15pF
47K
9
1
8
4
5
½6BM8
.003
50pF
1mH
RFC
22K
1w.
½6BM8
6
3
7
2
22K
.001
.01
1mH
RFC
1mF
10K
1w.
2.5mH
RFC
.01
.005
2kV
365pF
L1
L2
365pF
Plug in coils = 1¼" dia.
160 – L1 = 82T #20e.
L2 = 30T #28e.
80 – L1 = 40T #20e.
L2 = 15T #20e.
40 – L1 = 15T #20e.
L2 = 10T #25e.
L2 wind on bottom of L1 in the same direction
1K
KEY
1N4007
180K
Transformer
300-0-300a.c.
20ma
1N4007 (2)
1000 PIV
350
10w.
6.3v.
filament
40mF
400v.
40mF
400v.
1K
1w.
REC
TRANS
To receiver antenna input
Antenna
1a.
Fuse
Switch
120v.a.c.

Fig. 90-4

SIMPLE FM TRANSMITTER

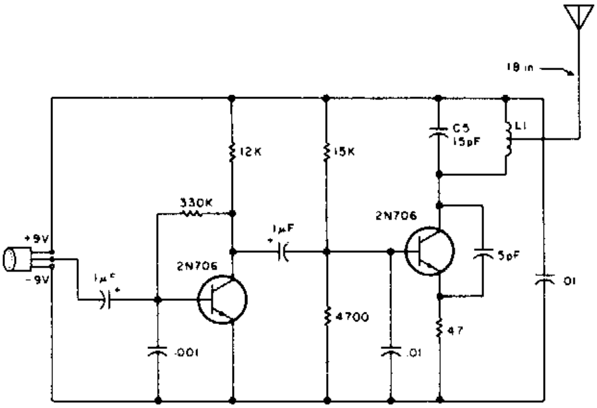

18 in
12K
15K
C5
15pF
L1
330K
2N706
1μF
2N706
5pF
+9V
-9V
1μF
.001
4700
.01
47
.01

Fig. 90-5

Circuit Notes

This transmitter can be tuned to the FM broadcast band, 2 meters,-or other VHF bands by changing C5 and L1. The values given for C5 and L1 will place the frequency somewhere in the FM broadcast band. L1 is 4 turns of #20 enameled wire airwound, ¼ inch in diameter, 5mm long and center-tapped. The microphone is an electret type and the antenna is 18 inches of any type of wire. Keep all leads as short as possible to minimize stray capacitance. The range of the transmitter is several hundred yards.

91

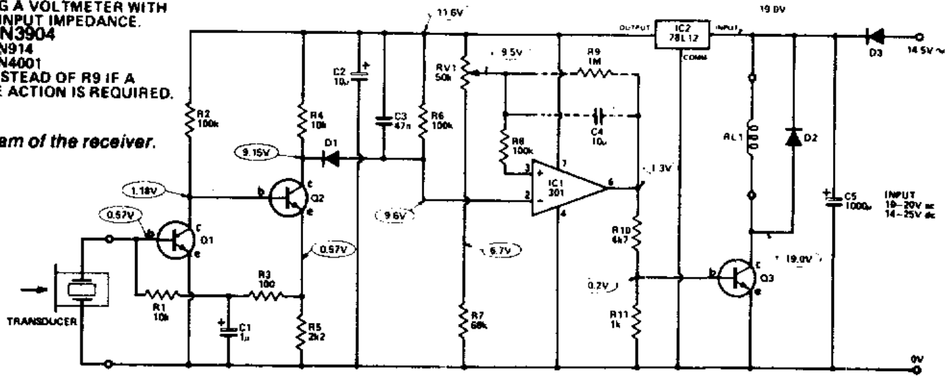
Ultrasonics

The sources of the following circuits are contained in the Sources section beginning on page 730. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

Ultrasonic Switch
Ultrasonic Bug-Chaser
Ultrasonic Pest Repeller
Mosquito-Repelling Circuit
40 kHz Ultrasonic Transmitter

ULTRASONIC SWITCH

NOTES:
VOLTAGES MEASURED WITH NO INPUT SIGNAL USING A VOLTMETER WITH 10 MEG OHM INPUT IMPEDANCE.
Q1–Q3 ARE 2N3904
D1 IS 1N914
D2,D3 ARE 1N4001
C4 IS USED INSTEAD OF R9 IF A MONOSTABLE ACTION IS REQUIRED.


11.6V
19.0V
IC2 78L12
OUTPUT
INPUT
COMM
D3
14.5V ~
R2 100k
R4 10k
C2 10µ
C3 47n
R6 100k
RV1 50k
9.5V
R9 1M
C4 10µ
R8 100k
IC1 301
1.3V
RL1
D2
C5 1000µ
INPUT 10–20V ac 14–25V dc
9.15V
D1
1.18V
Q2
0.57V
Q1
9.6V
6.7V
R10 4k7
0.2V
Q3
19.0V
TRANSDUCER
R1 10k
R3 100
C1 1µ
R5 2k2
R7 68k
R11 1k
0V

Circuit diagram of the receiver.

NOTE:
VOLTAGES MEASURED USING A VOLTMETER WITH 10 MEG OHM INPUT IMPEDANCE.

Q1-2 ARE 2N3904

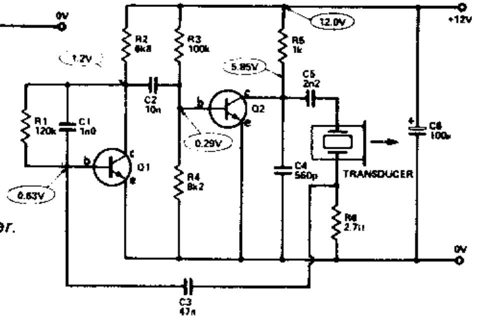

+12V
12.0V
R2 6k8
R3 100k
R5 1k
1.2V
5.85V
C5 2n2
C2 10n
Q2
R1 120k
C1 1n0
0.29V
C6 100µ
Q1
R4 8k2
C4 560p
TRANSDUCER
0.63V
R6 2.7Ω
0V
C3 47n

Fig. 91-1

Circuit diagram of the transmitter.

Circuit Notes

Receiver. Output from the transducer is amplified by Q1 and Q2, and rectified by D1. Voltage on pin 2 of IC1 will go more negative as the input signal increases. IC1 is used as a comparator and checks the voltage on pin 2 (i.e., the sound level), to that on pin 3 which is the reference level. If pin 2 is at a lower voltage than pin 3 (i.e., a signal is present), the output of IC1 will be high (about 10.5 volts) and this will turn on Q3 which will close the relay. The converse occurs if pin 2 is at a higher voltage than pine 3.

Transmitter. The oscillator frequency is determined by the transducer characteristics [(minimum (series resonance) at 39.8 kHz followed by a maximum (parallel resonance) at 41.5 kHz.)] Two transistors from a noninverting amplifier and positive feedback is supplied via the transducer, R6 and C3. At the series resonant frequency, this feedback is strong enough to cause oscillation.

ULTRASONIC BUG-CHASER

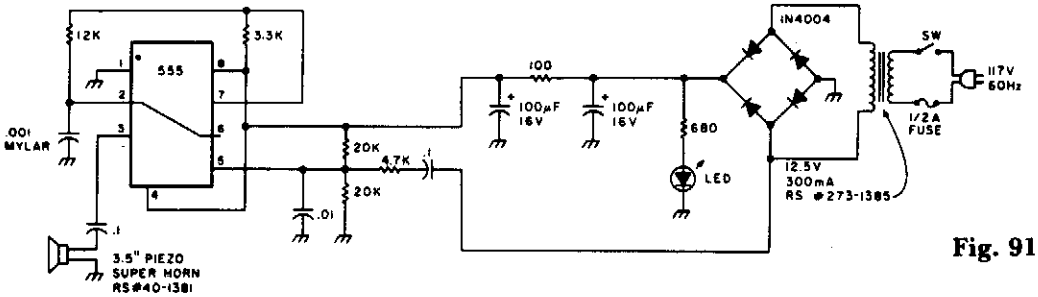

12K
3.3K
555
1
2
3
4
5
6
7
8
.001
MYLAR
.1
3.5" PIEZO
SUPER HORN
RS#40-1381
20K
20K
.01
4.7K
.1
100
100μF
16V
100μF
16V
680
LED
IN4004
12.5V
300mA
RS #273-1385
SW
117V
60Hz
1/2A
FUSE

Fig. 91-2

Circuit Notes

Low-intensity ultrasonic sound waves in the 30-45 kHz frequency band repel insects and small rodents. The unit is designed to generate a swept square wave from 30 to 45 kHz. The LM555 IC is wired as an ultrasonic oscillator driving a piezoelectric speaker of the hi-fi super-tweeter type. The output of the oscillator is swept by a 60-Hz signal from the ac input of the bridge rectifier. The LED acts as a pilot.

MOSQUITO-REPELLING CIRCUIT

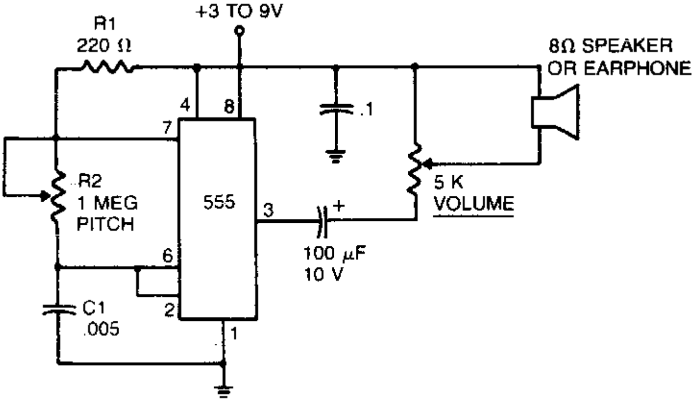

R1
220 Ω
+3 TO 9V
4
8
7
555
3
6
2
1
R2
1 MEG
PITCH
C1
.005
.1
100 μF
10 V
+
5 K
VOLUME
8Ω SPEAKER
OR EARPHONE

Fig. 91-3

Circuit Notes

In the 555 oscillator circuit, adjusting R2 will provide output frequencies from below 200 Hz to above 62 kHz. Use a good quality miniature speaker so that it will produce frequencies on the order of 20 kHz.

ULTRASONIC PEST REPELLER

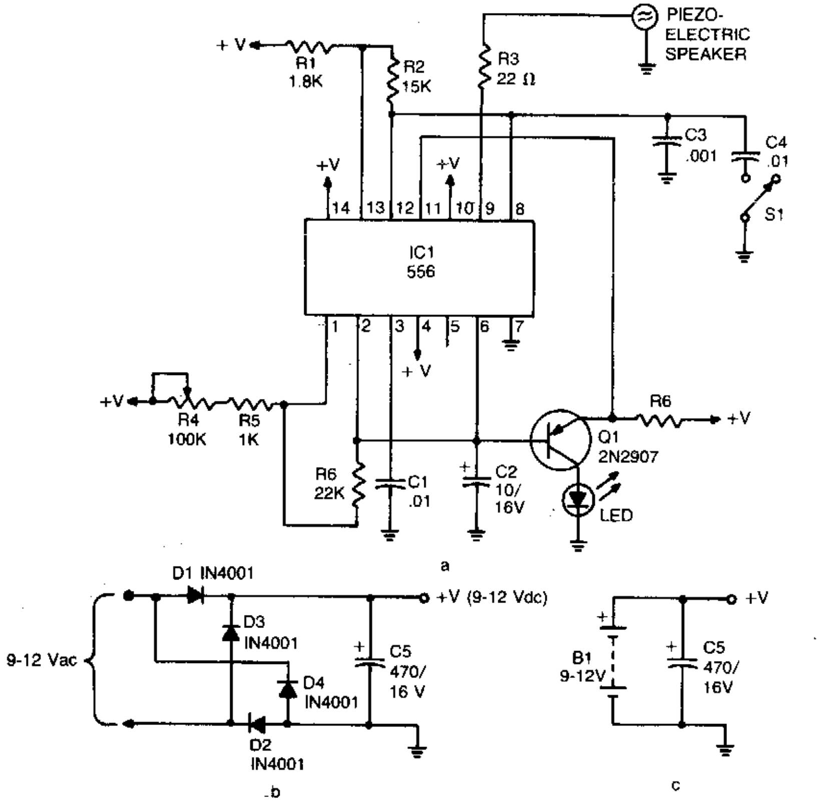

PIEZO-
ELECTRIC
SPEAKER
+V
R1
1.8K
R2
15K
R3
22 Ω
C3
.001
C4
.01
S1
+V
+V
14
13
12
11
10
9
8
IC1
556
1
2
3
4
5
6
7
+ V
+V
R4
100K
R5
1K
R6
+V
Q1
2N2907
R6
22K
C1
.01
C2
10/
16V
LED
a
D1 IN4001
+V (9-12 Vdc)
D3
IN4001
C5
470/
16 V
9-12 Vac
D4
IN4001
D2
IN4001
b
+V
B1
9-12V
C5
470/
16V
c

Fig. 91-4

Circuit Notes

The device emits ultrasonic sound waves that sweep between 65,000 and 25,000 hertz. Designed around a 556 dual timer, one half operated as an astable multivibrator with an adjustable frequency of 1 to 3 Hz. The second half is also operated as an astable multivibrator but with a fixed free running frequency around 45,000 Hz. The 25-65 kHz sweep is accomplished by coupling the voltage across C2 (the timing capacitor for the first half of the 556) via Q1 to the control voltage terminal (pin 11) of the second half of the 556. The device that radiates the ultrasonic sound is a piezo tweeter.

40 kHz ULTRASONIC TRANSMITTER

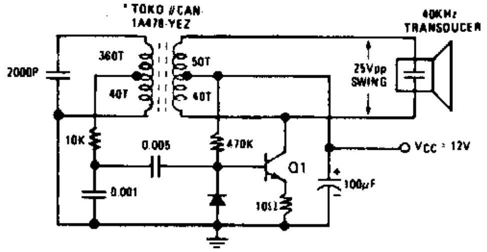

* TOKO #CAN
1A478-YEZ
40KHz
TRANSDUCER
360T
40T
50T
40T
2000P
25Vpp
SWING
10K
0.005
470K
Q1
Vcc = 12V
0.001
10Ω
100μF

Fig. 91-5

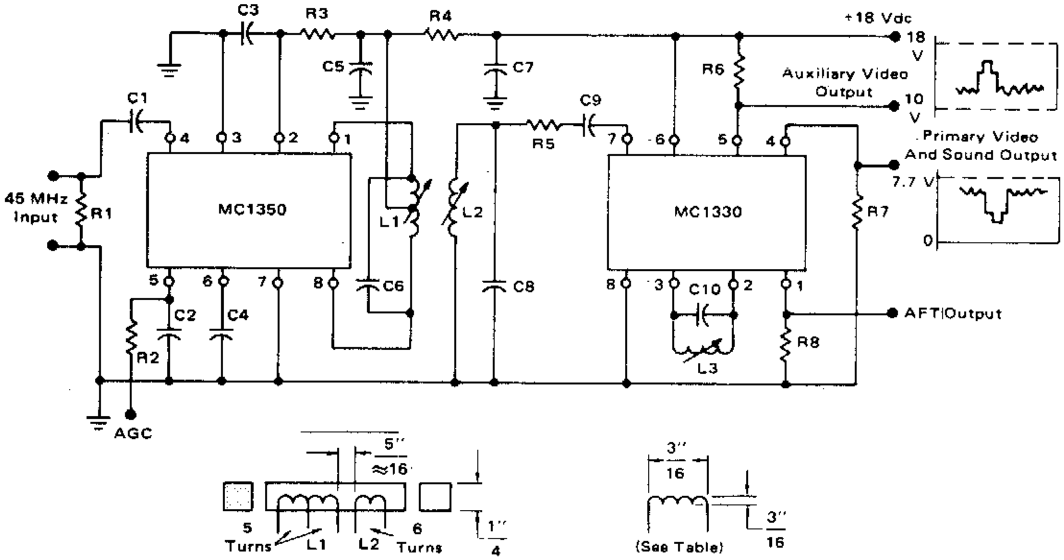
92

Video Amplifiers

The sources of the following circuits are contained in the Sources section beginning on page 730. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

Video IF Amplifier and Low-Level Video Detector Circuit
Television IF Amplifier and Detector Using an MC1330 and an MC1352
Two-Stage Wideband Amplifier
Video IF Amplifier and Low-Level Video Detector Circuit
TV Sound IF or FM IF Amplifier with Quadrature Detector
IF Amplifier
FET Cascode Video Amplifier
High Impedance Low Capacitance Amplifier
JFET Bipolar Cascode Video Amplifier
Video Amplifier
Video Amplifier

VIDEO IF AMPLIFIER AND LOW-LEVEL VIDEO DETECTOR CIRCUIT


C3
R3
R4
+18 Vdc
18 V
C5
C7
R6
Auxiliary Video Output
10 V
C1
C9
R5
4
3
2
1
7
6
5
4
Primary Video And Sound Output
7.7 V
0
45 MHz Input
R1
MC1350
L1
L2
MC1330
R7
5
6
7
8
C6
C8
8
3
C10
2
1
AFT Output
C2
C4
R2
R8
L3
AGC
5"/16
1"/4
5 Turns
L1
L2
6 Turns
3"/16
16
(See Table)
3"/16

All windings #30 AWG tinned nylon acetate wire tuned with high permiability slugs. Coil Craft #4786 differential transformer.

L3 wound with #26 AWG tinned nylon acetate wire tuned by distorting winding.

C1 = 0.001 µF
C2 = 0.002 µF
C3 = 0.002 µF
C4 = 0.002 µF
C5 = 0.002 µF

C6 = See Table
C7 = 0.1 µF
C8 = See Table
C9 = 68 pF
C10 = See Table

R1 = 50 Ω
R2 = 5 k
R3 = 470 Ω
R4 = 220 Ω
R5 = 22 Ω

R6 = 3.3 kΩ
R7 = 3.9 kΩ
R8 = 3.9 kΩ
All Resistors 1/4-W ±10%

All Caps Marked µF Ceramic HiK
All Caps Marked pF Silver Mica 5%

Table of Component Values

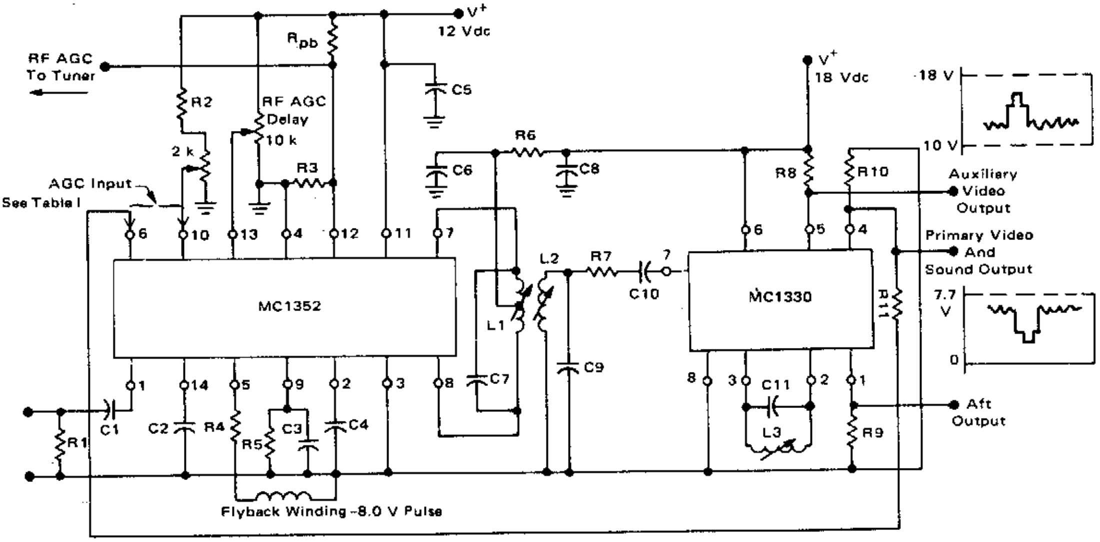
Component	36 MHz	45 MHz	58 MHz
C6	24 pF	15 pF	10 pF
C8	18 pF	12 pF	10 pF
C10	33 pF	33 pF	18 pF
L3	12 Turns	10 Turns	10 Turns

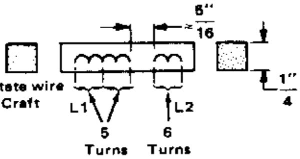
Fig. 92-1

Circuit Notes

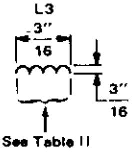
The circuit has a typical voltage gain of 84 dB and a typical AGC range of 80 dB. It gives very small changes in bandpass shape, usually less than 1 dB tilt for 60 dB compression. There are no shielded sections. The detector uses a single tuned circuit (L3 and C10). Coupling between the two integrated circuits is achieved by a double tuned transformer (L1 and L2).

TELEVISION IF AMPLIFIER AND DETECTOR USING AN MC1330 AND AN MC1352


V+
12 Vdc
RF AGC
To Tuner
Rpb
C5
R2
RF AGC
Delay
10 k
2 k
R3
AGC Input
See Table I
6
10
13
4
12
11
7
MC1352
1
14
5
9
2
3
8
R6
C6
C8
L2
R7
C10
7
L1
C7
C9
C1
C2
R4
R5
C3
C4
R1
Flyback Winding –8.0 V Pulse
V+
18 Vdc
18 V
10 V
R8
R10
Auxiliary
Video
Output
6
5
4
Primary Video
And
Sound Output
MC1330
R11
7.7
V
0
8
3
C11
2
1
L3
R9
Aft
Output


5″
16
1″
4
L1
L2
5
Turns
6
Turns

All windings #30 AWG tinned nylon acetate wire tuned with high permiability slugs. Coil Craft #4786 differential transformer.


L3
3″
16
3″
16
See Table II

Wound with #26 AWG tinned nylon acetate wire tuned by distorting winding.

C10 = 62 pF
C11 = (See Table II)
All Resistors 1/4-Watt ±5%

TABLE I

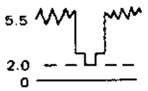
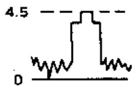
Video Polarity	Pin 6 Voltage	Pin 10 Voltage	R4
Negative-Going Sync.		Adj. 1.0–4.0 Vdc Nom 2.0 V	0
Positive-Going Sync.	Adj. 1.0–8.0 Vdc Nom 4.5 V		3.9 k

TABLE II

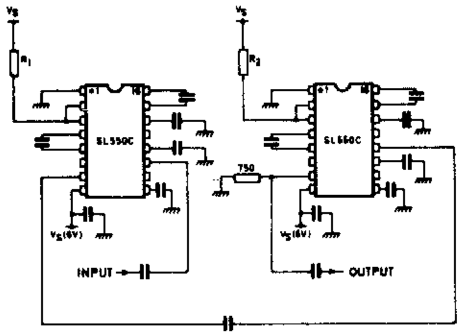
Component	36 MHz	45 MHz	58 MHz
C7	24 pF	15 pF	10 pF
C9	18 pF	12 pF	10 pF
C11	33 pF	33 pF	18 pF
L3	12 Turns	10 Turns	10 Turns

Rpb (See Text)
R1 = 50 Ω
R2 = 3.9 kΩ
R3 = (See Text)
R4 = (See Table I)
R5 = 220 kΩ
R6 = 220 Ω
R7 = 22 Ω
R8 = 3.3 kΩ
R9 = 3.9 kΩ

R10 = 3.9 kΩ
R11 = 4.7 kΩ
C1 = 0.001 μF
C2 = 0.1 μF
C3 = 0.25 μF
C5 = 0.1 μF
C6 = 0.1 μF
C7 = (See Table II)
C8 = 0.1 μF
C9 = (See Table II)

Fig. 92-2

TWO-STAGE WIDEBAND AMPLIFIER



Vs
R1
SL550C
Vs(6V)
INPUT
Vs
R2
SL550C
750
Vs(6V)
OUTPUT

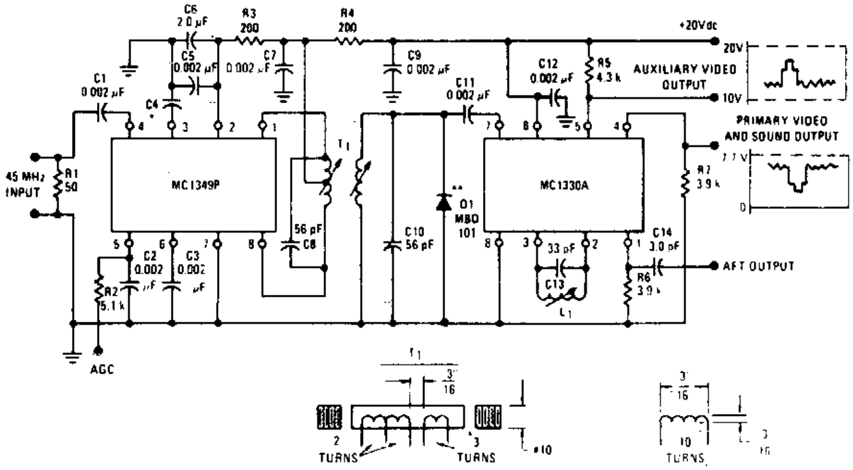
ALL CAPACITORS 1000 pF

Fig 92-3

Circuit Notes

A wideband high gain configuration using two SL550s connected in series. The first stage is connected in common emitter configuration, the second stage is a common base circuit. Stable gains of up to 65 dB can be achieved by the proper choice of R1 and R2. The bandwidth is 5 to 130 MHz, with a noise figure only marginally greater than the 2.0 dB specified for a single stage circuit.

VIDEO IF AMPLIFIER AND LOW-LEVEL VIDEO DETECTOR CIRCUIT



C6
2.0 µF
R3
200
R4
200
+20Vdc
C5
0.002 µF
C7
0.002 µF
C9
0.002 µF
C12
0.002 µF
R5
4.3 k
AUXILIARY VIDEO OUTPUT
20V
10V
C1
0.002 µF
C4
C11
0.002 µF
PRIMARY VIDEO AND SOUND OUTPUT
7.7 V
0
45 MHz INPUT
R1
50
MC1349P
T1
D1
MBD
101
MC1330A
R7
3.9 k
56 pF
C8
C10
56 pF
C14
3.0 pF
C2
0.002 µF
C3
0.002 µF
33 pF
C13
L1
AFT OUTPUT
R6
3.9 k
R2
5.1 k
AGC
T1
3/16
2 TURNS
3 TURNS
#10
3/16
10 TURNS
3/16

All windings 22 AWG tinned nylon acetate wire tuned with Coilcraft #61 slugs, size 10-32, or equivalent

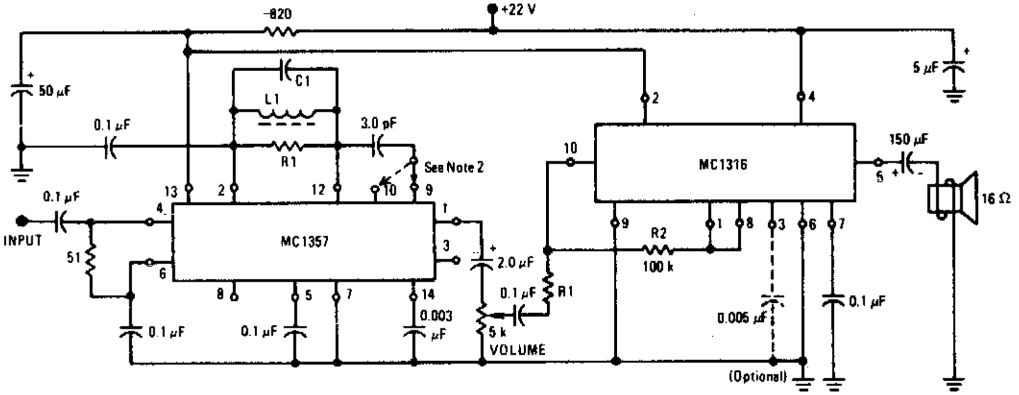
L1 wound with 26 AWG tinned nylon acetate wire tuned by distorting winding

*See Note 1 (page 3), and C4, Parts List (page 4) for this specification on the MC1349P Data Sheet

**See Input Overload Section of the Design Characteristics Page 3, and General Information, Page 5 Note 6

Fig. 92-4

TV SOUND IF OR FM IF AMPLIFIER WITH QUADRATURE DETECTOR


+22 V
820
50 µF
0.1 µF
C1
L1
R1
3.0 pF
See Note 2
13
2
12
10
9
0.1 µF
INPUT
51
4
6
MC1357
1
3
8
5
7
14
0.1 µF
0.1 µF
0.003 µF
2.0 µF
0.1 µF
R1
5 k
VOLUME
MC1316
2
4
10
9
R2
100 k
1
8
3
6
7
0.005 µF
(Optional)
0.1 µF
5 µF
150 µF
5
16 Ω

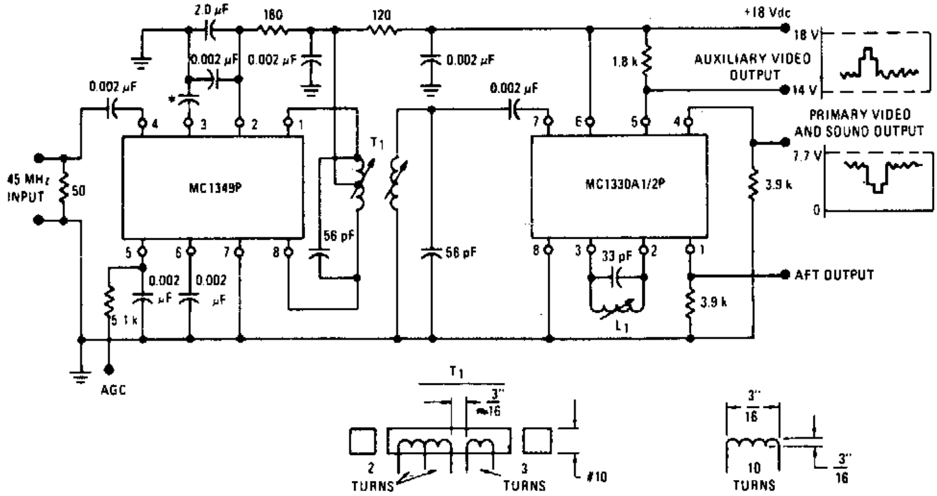
Typical Performance:
2 Watts Output
2% Distortion
250 µV Sensitivity (3 dB Lim.)

C1 = 120 pF
L1 = 14 µH
R1 = 20 kΩ
Q = 30

Fig. 92-5

IF AMPLIFIER

– TYPICAL APPLICATION OF MC1349P VIDEO IF AMPLIFIER and MC1330 LOW-LEVEL VIDEO DETECTOR CIRCUIT


2.0 µF
160
120
+18 Vdc
0.002 µF
0.002 µF
0.002 µF
0.002 µF
*
45 MHz INPUT
50
4
3
2
1
MC1349P
T1
56 pF
5
6
7
8
0.002 µF
0.002 µF
5.1 k
AGC
56 pF
0.002 µF
7
6
5
4
MC1330A1/2P
8
3
2
1
33 pF
L1
1.8 k
AUXILIARY VIDEO OUTPUT
18 V
14 V
PRIMARY VIDEO AND SOUND OUTPUT
7.7 V
0
3.9 k
3.9 k
AFT OUTPUT
T1
3/16"
2 TURNS
3 TURNS
#10
3/16"
10 TURNS
3/16"

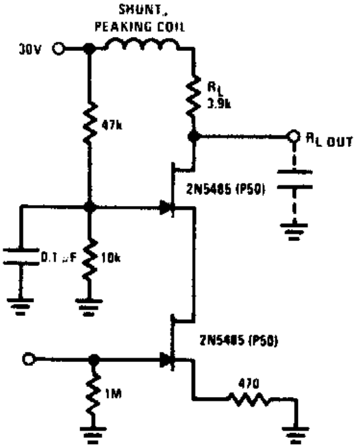
All windings #22 AWG tinned nylon acetate wire tuned with Coilcraft #61 slugs, size 10-32, or equivalent.

L1 wound with #26 AWG tinned nylon acetate wire tuned by distorting winding.

*See Note 1 (page 3), and C4, Parts List (page 4) of this specification.

Fig. 92-6

FET CASCODE VIDEO AMPLIFIER


SHUNT,
PEAKING COIL
30V
R_L
3.9k
47k
R_L OUT
2N5485 (P50)
0.1 μF
10k
2N5485 (P50)
470
1M

Circuit Notes

The FET cascode video amplifier features very low input loading and reduction of feedback to almost zero. The 2N5485 is used because of its low capacitance and high Y_{fs}. Bandwidth of this amplifier is limited by RL and load capacitance.

Fig. 92-7

HIGH IMPEDANCE LOW CAPACITANCE AMPLIFIER

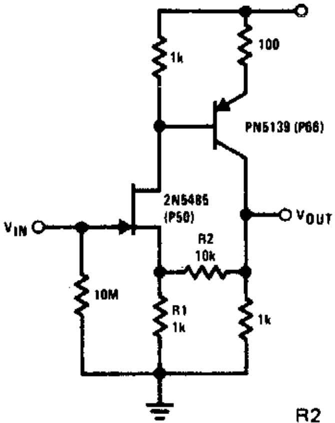

100
1k
PN5139 (P66)
2N5485
(P50)
V_{IN}
V_{OUT}
R2
10k
10M
R1
1k
1k

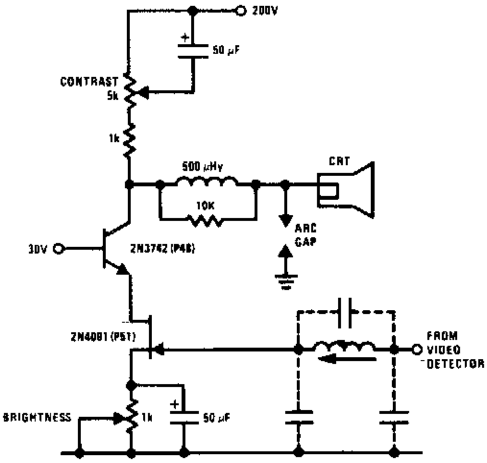
Fig. 92-8

$$V_{OUT} \simeq \frac{R2}{R1} V_{IN}$$

Circuit Notes

This compound series-feedback circuit provides high input impedance and stable, wide-band gain for general purpose video amplifier applications.

JFET BIPOLAR CASCODE VIDEO AMPLIFIER


200V
50 μF
CONTRAST
5k
1k
500 μHy
CRT
10K
ARC
GAP
30V
2N3742 (P48)
2N4091 (P51)
FROM
VIDEO
DETECTOR
BRIGHTNESS
1k
50 μF

Circuit Notes

The JFET-bipolar cascode circuit will provide full video output for the CRT cathode drive. Gain is about 90. The cascode configuration eliminates Miller capacitance problems with the 2N4091 JFET, thus allowing direct drive from the video detector. An m-derived filter using stray capacitance and a variable inductor prevents 4.5 MHz sound frequency from being amplified by the video amplifier.

Fig. 92-9

VIDEO AMPLIFIER

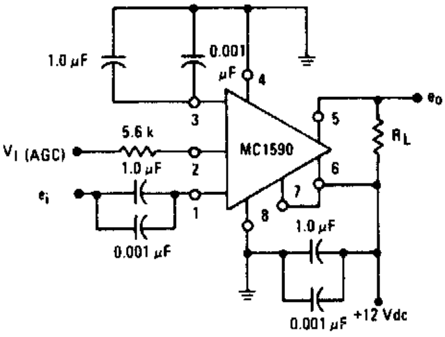
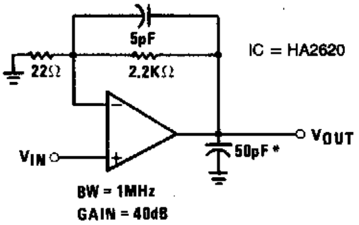

1.0 μF
0.001 μF
4
3
eo
5
RL
5.6 k
VI (AGC)
MC1590
1.0 μF
2
6
ei
1
7
8
1.0 μF
0.001 μF
0.001 μF
+12 Vdc

Fig. 92-10

VIDEO AMPLIFIER


5pF
22Ω
2.2KΩ
IC = HA2620
VIN
VOUT
50pF *
BW = 1MHz
GAIN = 40dB

*A small load capacitance of at least 30pF (including stray capacitance) is recommended to prevent possible high frequency oscillations.

Fig. 92-11

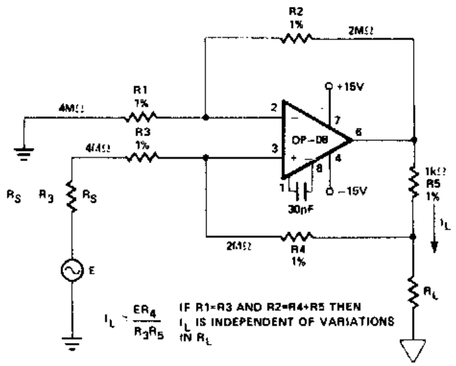
93

Voltage and Current Sources and References

The sources of the following circuits are contained in the Sources section beginning on page 730. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

Bilateral Current Source
0 V to 20 V Power Reference
Programmable Voltage Source
Bilateral Current Source
Noninverting Bipolar Current Source
Voltage Reference
Low Voltage Adjustable Reference Supply
Voltage Reference
Low Power Regulator Reference
High Stability Voltage Reference
± 3 V Reference
± 5 V Reference
Zenerless Precision Millivolt Source
± 10 V Reference
Precision Reference Square Wave Voltage Reference
Inverting Bipolar Current Source
Precision Reference Micropower 10 V Reference
Precision Reference Low Noise Buffered Reference
Constant Current Source
Precision Dual Tracking Voltage References
Precision Reference Bipolar Output Reference
Precision Reference 0 V to 20 V Power Reference
Precision Reference Standard Cell Replacement

BILATERAL CURRENT SOURCE


R2
1%
2MΩ
R1
1%
4MΩ
R3
1%
4MΩ
+15V
2
3
7
6
4
8
1
OP-08
−15V
30nF
1kΩ
R5
1%
I_L
R_S
R_3
R_S
2MΩ
R4
1%
E
R_L

$I_L = \frac{ER_4}{R_3R_5}$ IF R1=R3 AND R2=R4+R5 THEN I_L IS INDEPENDENT OF VARIATIONS IN R_L

Circuit Notes

The circuit will produce the current relationship to within 2% using 1% values for R1 through R5. This includes variations in R_L from 100 ohm to 2000 ohm. The use of large resistors for R1 through R4 minimizes the error due to R_L variations. The large resistors are possible because of the excellent input bias current performance of the OP-08.

Fig. 93-1

0 V TO 20 V POWER REFERENCE

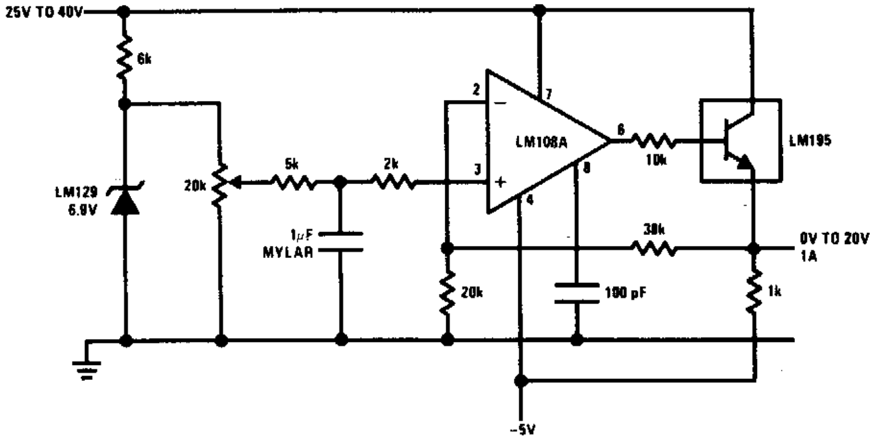
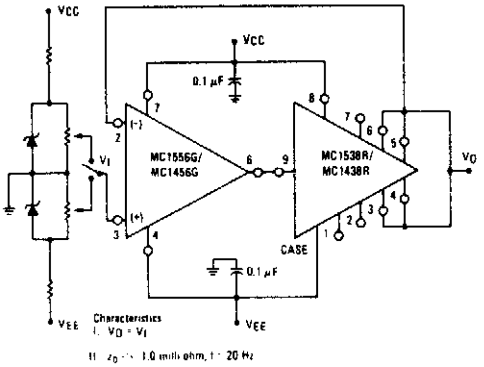

25V TO 40V
6k
LM129
6.9V
20k
5k
2k
1µF
MYLAR
2
3
7
6
8
4
LM108A
10k
LM195
30k
0V TO 20V
1A
20k
100 pF
1k
−5V

Fig. 93-2

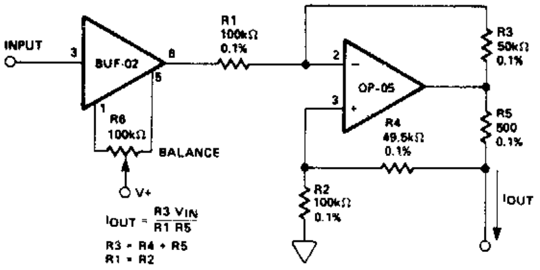
PROGRAMMABLE VOLTAGE SOURCE


VCC
VCC
0.1 µF
V_I
(−)
(+)
2
3
7
4
6
9
MC1556G/
MC1456G
MC1538R/
MC1438R
8
7
6
5
4
3
2
1
CASE
0.1 µF
V_O
VEE
VEE

Characteristics
I. $V_O = V_I$
II. $z_O \leq$ 1.0 milli ohm, f ≤ 20 Hz

Fig. 93-3

BILATERAL CURRENT SOURCE


INPUT
3
BUF-02
6
5
1
R6
100kΩ
BALANCE
V+
R1
100kΩ
0.1%
2
3
OP-05
R4
49.5kΩ
0.1%
R2
100kΩ
0.1%
R3
50kΩ
0.1%
R5
500
0.1%
IOUT

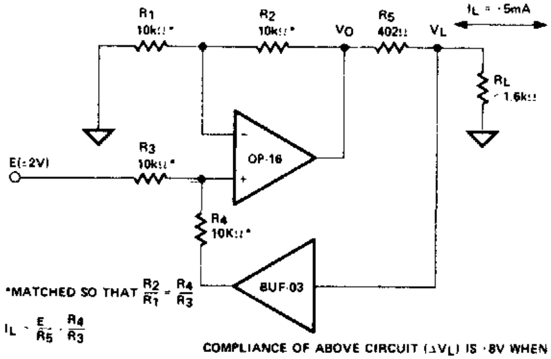
$I_{OUT} = \frac{R3\ V_{IN}}{R1\ R5}$

$R3 = R4 + R5$

$R1 = R2$

Fig. 93-4

NONINVERTING BIPOLAR CURRENT SOURCE


R1
10kΩ*
R2
10kΩ*
VO
R5
402Ω
VL
IL = ±5mA
RL
≤1.6kΩ
R3
10kΩ*
E(±2V)
OP-16
R4
10KΩ*
BUF-03

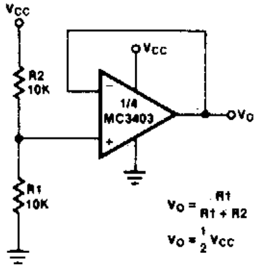
*MATCHED SO THAT $\frac{R2}{R1} = \frac{R4}{R3}$

$I_L = \frac{E}{R5} \cdot \frac{R4}{R3}$

Fig. 93-5

COMPLIANCE OF ABOVE CIRCUIT (ΔV_L) IS ±8V WHEN E = ±2V AND $R_L \leq$ 1.6kΩ. NOTE THAT V_O IS ±10V UNDER THESE CONDITIONS.

VOLTAGE REFERENCE

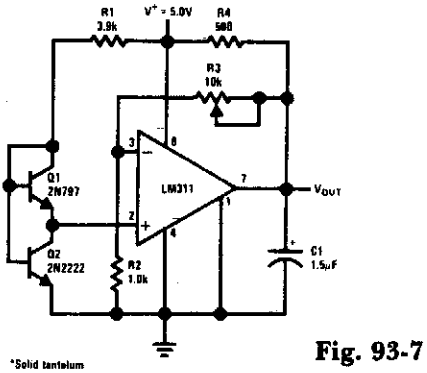

VCC
R2
10K
VCC
1/4
MC3403
VO
R1
10K

$V_O = \frac{R1}{R1 + R2}$

$V_O = \frac{1}{2} V_{CC}$

Fig. 93-6

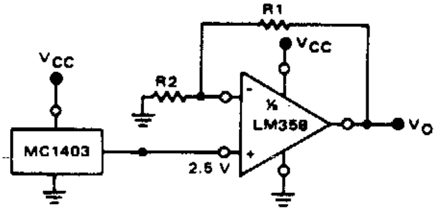
LOW VOLTAGE ADJUSTABLE REFERENCE SUPPLY


R1
3.9k
V+ = 5.0V
R4
500
R3
10k
Q1
2N797
Q2
2N2222
R2
1.0k
3
2
8
4
7
1
LM311
VOUT
C1
1.5µF

*Solid tantalum

Fig. 93-7

VOLTAGE REFERENCE


R1
VCC
VCC
R2
½
LM358
VO
MC1403
2.5 V

$V_O = 2.5\ V(1 + \frac{R1}{R2})$

Fig. 93-8

LOW POWER REGULATOR REFERENCE

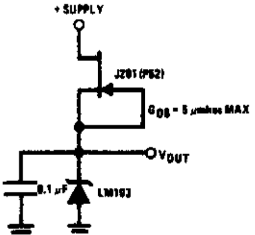

+SUPPLY
J201 (P52)
GOS = 5 µmhos MAX
VOUT
0.1 µF
LM193

Fig. 93-9

Circuit Notes

This simple reference circuit provides a stable voltage reference almost totally free of supply voltage hash. Typical power supply rejection exceeds 100 dB.

HIGH STABILITY VOLTAGE REFERENCE

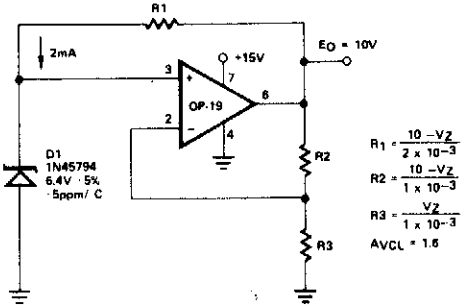

R1
2mA
EO = 10V
+15V
3
7
OP-19
6
2
4
D1
1N45794
6.4V · 5%
·5ppm/ C
R2
R3
R1 = (10 −VZ)/(2 x 10−3)
R2 = (10 −VZ)/(1 x 10−3)
R3 = VZ/(1 x 10−3)
AVCL = 1.6

Fig. 93-10

ZENERLESS PRECISION MILLIVOLT SOURCE

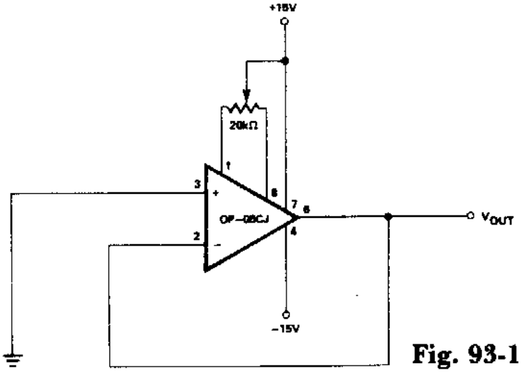

+15V
20kΩ
1
3
8
7
6
OP-08CJ
2
4
−15V
VOUT

Fig. 93-13

± 3 V REFERENCE

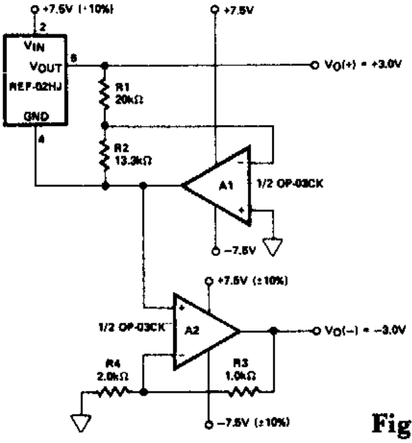

+7.5V (±10%)
2
VIN
VOUT
REF-02HJ
GND
6
4
R1
20kΩ
R2
13.3kΩ
+7.5V
VO(+) = +3.0V
A1
1/2 OP-03CK
−7.5V
+7.5V (±10%)
1/2 OP-03CK
A2
VO(−) = −3.0V
R4
2.0kΩ
R3
1.0kΩ
−7.5V (±10%)

Fig. 93-11

± 10 V REFERENCE

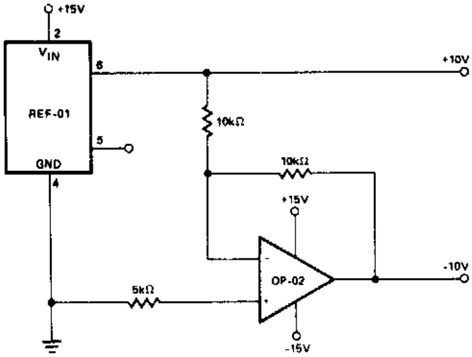

+15V
2
VIN
6
REF-01
5
GND
4
+10V
10kΩ
10kΩ
+15V
OP-02
5kΩ
-15V
-10V

Fig. 93-14

± 5 V REFERENCE

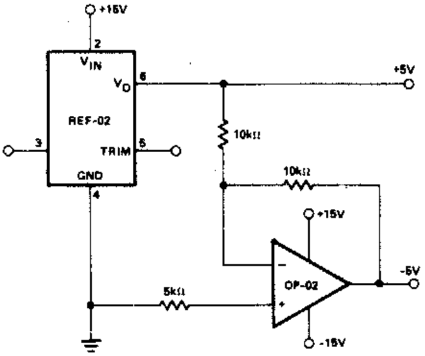

+15V
2
VIN
VO
6
REF-02
3
TRIM
5
GND
4
+5V
10kΩ
10kΩ
+15V
OP-02
5kΩ
-15V
-5V

Fig. 93-12

PRECISION REFERENCE SQUARE WAVE VOLTAGE REFERENCE

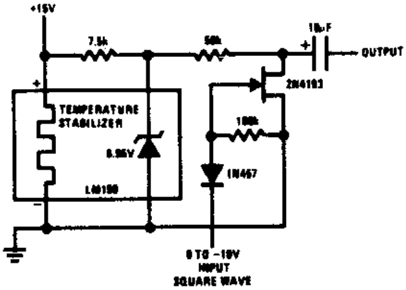
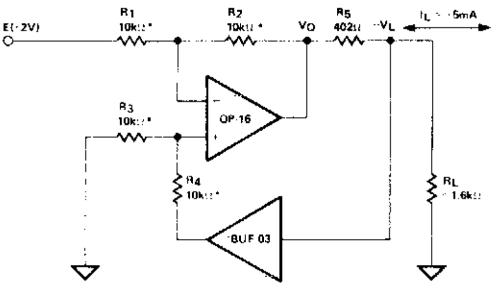

+15V
7.5k
50k
10µF
OUTPUT
2N4193
TEMPERATURE
STABILIZER
6.95V
LM199
100k
1N457
0 TO −15V
INPUT
SQUARE WAVE

Fig. 93-15

INVERTING BIPOLAR CURRENT SOURCE (HIGH SPEED)


E(−2V)
R1
10kΩ*
R2
10kΩ*
VO
R5
402Ω
−VL
IL = −5mA
R3
10kΩ*
OP-16
R4
10kΩ*
BUF 03
RL
1.6kΩ

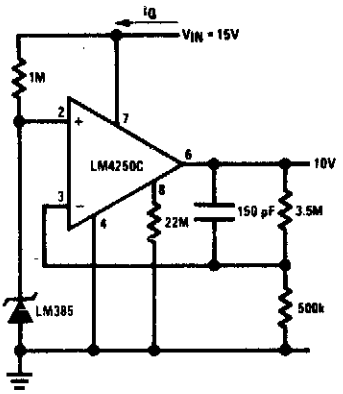
*MATCHED SO THAT $\frac{R_2}{R_1} = \frac{R_4}{R_3}$

$I_L = -\frac{E}{R_5} \cdot \frac{R_2}{R_1}$

COMPLIANCE OF ABOVE CIRCUIT (ΔVL) IS ±8V WHEN E = −2V AND RL = 1.6kΩ. NOTE THAT VO is +10V UNDER THESE CONDITIONS.

Fig. 93-16

PRECISION REFERENCE MICROPOWER 10 V REFERENCE


IQ
VIN = 15V
1M
2
7
LM4250C
6
10V
3
8
4
22M
150 pF
3.5M
LM385
500k

*IQ≥20 μA standby current

Fig. 93-17

PRECISION REFERENCE LOW NOISE BUFFERED REFERENCE

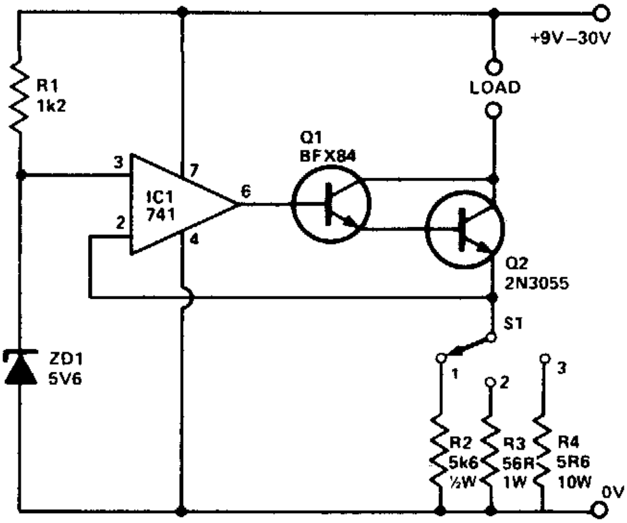

+9V–30V
LOAD
R1
1k2
Q1
BFX84
3
7
IC1
741
6
2
4
Q2
2N3055
S1
ZD1
5V6
1
2
3
R2
5k6
½W
R3
56R
1W
R4
5R6
10W
0V

Fig. 93-18

Circuit Notes

The circuit will provide 3 preset currents which will remain constant despite variations of ambient temperature or line voltage. ZD1 produces a temperature stable reference voltage which is applied to the noninverting input of IC1. 100% feedback is applied from the output to the inverting input holding the voltage at Q2s emitter at the same potential as the noninverting input. The current flowing into the load therefore is defined solely by the resistor selected by S1. With the values employed here, a preset current of 10 mA, 100 mA or 1 A can be selected. Q2 should be mounted on a suitable heatsink.

CONSTANT CURRENT SOURCE

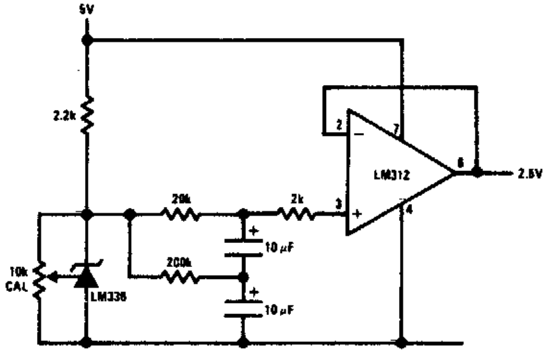

5V
2.2k
20k
2k
200k
10 μF
10 μF
10k
CAL
LM336
2
7
3
6
4
LM312
2.5V

Fig. 93-19

PRECISION DUAL TRACKING VOLTAGE REFERENCES

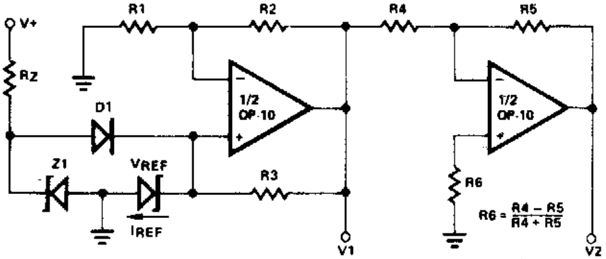

V+
RZ
R1
R2
R4
R5
D1
Z1
VREF
IREF
R3
1/2
OP-10
1/2
OP-10
R6
R6 = $\frac{R4 - R5}{R4 + R5}$
V1
V2

Fig. 93-20

PRECISION REFERENCE BIPOLAR OUTPUT REFERENCE

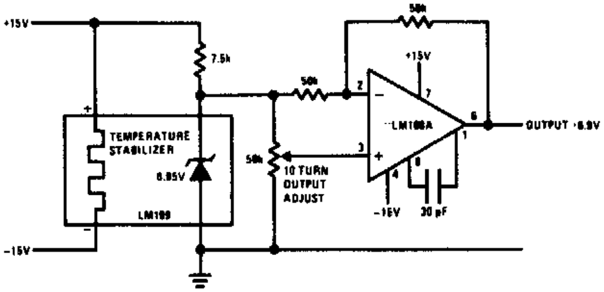

+15V
-15V
7.5k
50k
50k
+15V
2
7
6
3
1
4
8
LM108A
OUTPUT ·6.9V
TEMPERATURE
STABILIZER
6.95V
LM199
50k
10 TURN
OUTPUT
ADJUST
-15V
30 pF

Fig. 93-21

PRECISION REFERENCE 0 V TO 20 V POWER REFERENCE

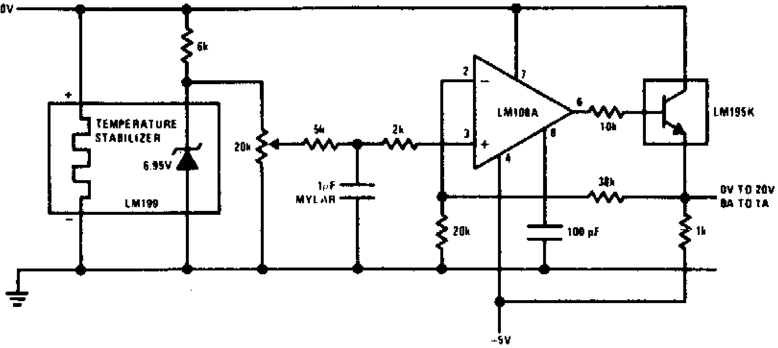

25V TO 40V
6k
+
TEMPERATURE STABILIZER
6.95V
LM199
−
20k
5k
2k
1μF MYLAR
2
3
−
+
LM108A
7
6
8
4
10k
LM195K
38k
0V TO 20V 0A TO 1A
20k
100 pF
1k
−5V

Fig. 93-22

PRECISION REFERENCE STANDARD CELL REPLACEMENT

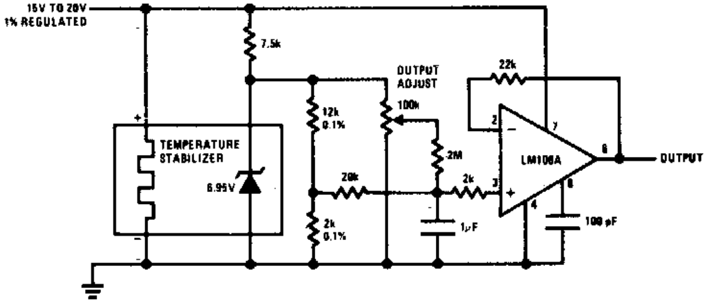

15V TO 20V 1% REGULATED
7.5k
+
TEMPERATURE STABILIZER
6.95V
−
12k 0.1%
2k 0.1%
20k
OUTPUT ADJUST
100k
2M
2k
1μF
22k
2
3
−
+
LM108A
7
6
8
4
100 pF
OUTPUT

Fig. 93-23

94

Voltage-Controlled Oscillators

The sources of the following circuits are contained in the Sources section beginning on page 730. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

Linear Voltage Controlled Oscillator
10 Hz to 10 kHz Voltage Controlled Oscillator
Precision Voltage Controlled Oscillator
Voltage Controlled Oscillator
Simple Voltage Controlled Oscillator
Three Decades VCO
Two-Decade High-Frequency VCO
Voltage Controlled Oscillator
Voltage Controlled Oscillator

LINEAR VOLTAGE CONTROLLED OSCILLATOR

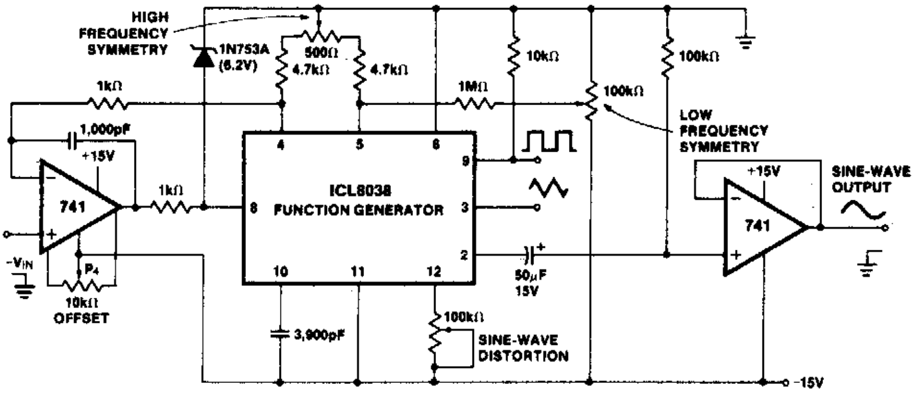
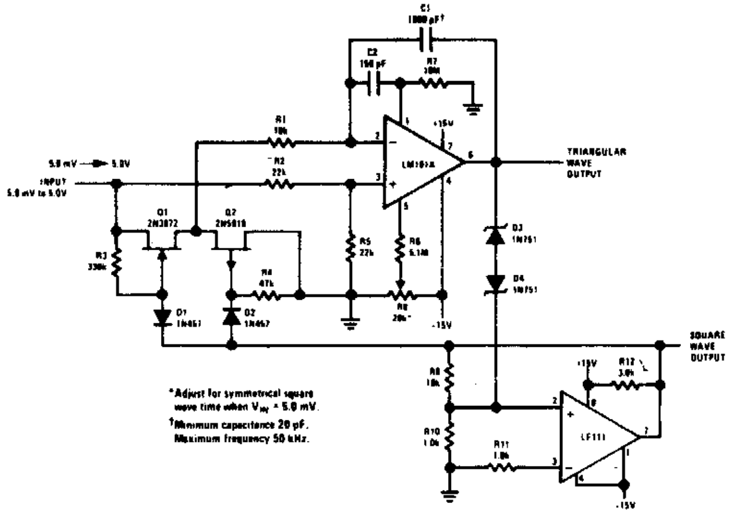

HIGH FREQUENCY SYMMETRY
1N753A (6.2V)
500Ω
4.7kΩ
4.7kΩ
10kΩ
100kΩ
1kΩ
1MΩ
100kΩ
LOW FREQUENCY SYMMETRY
1,000pF
+15V
741
1kΩ
4
5
6
9
8
3
2
10
11
12
ICL8038 FUNCTION GENERATOR
+15V
SINE-WAVE OUTPUT
741
-VIN
P4
10kΩ OFFSET
50μF 15V
3,900pF
100kΩ
SINE-WAVE DISTORTION
-15V

Fig. 94-1

Circuit Notes

The linearity of input sweep voltage versus output frequency is significantly improved by using an op amp.

10 Hz TO 10 kHz VOLTAGE CONTROLLED OSCILLATOR

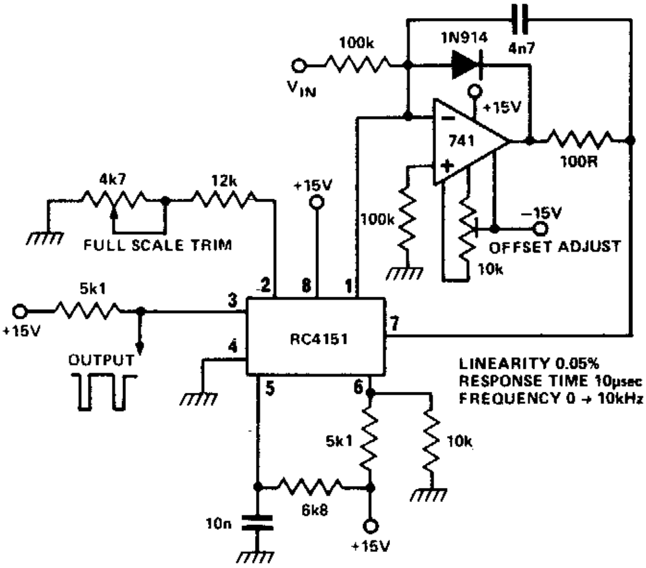

C1 1000 pF†
C2 150 pF
R7 10M
R1 10k
R2 22k
+15V
LM101A
1
2
3
4
5
6
7
5.0 mV → 5.0V
INPUT 5.0 mV to 5.0V
TRIANGULAR WAVE OUTPUT
Q1 2N3972
Q2 2N5019
R3 330k
R4 47k
R5 22k
R6 5.1M
R8 20k*
-15V
D1 1N457
D2 1N457
D3 1N751
D4 1N751
SQUARE WAVE OUTPUT
R9 10k
R10 1.0k
R11 1.0k
R12 3.0k
+15V
LF111
-15V
2
3
4
7
8
1

*Adjust for symmetrical square wave time when V_{IN} = 5.0 mV.

†Minimum capacitance 20 pF. Maximum frequency 50 kHz.

Fig. 94-2

PRECISION VOLTAGE CONTROLLED OSCILLATOR



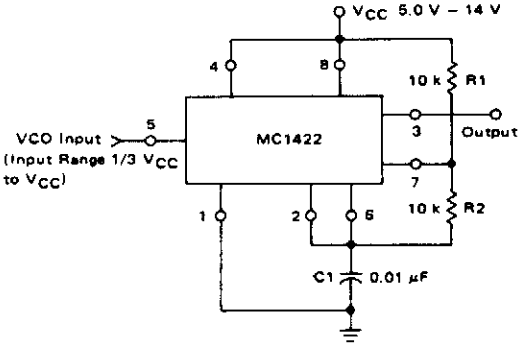
100k
1N914
4n7
V_IN
+15V
741
100R
4k7
12k
+15V
FULL SCALE TRIM
100k
−15V
OFFSET ADJUST
10k
5k1
3
2
8
1
+15V
RC4151
7
4
OUTPUT
5
6
LINEARITY 0.05%
RESPONSE TIME 10μsec
FREQUENCY 0 → 10kHz
5k1
10k
6k8
10n
+15V

Fig. 94-3

Circuit Notes

RC 4151 precision voltage-to-frequency converter generates a pulse train output linearly proportional to the input voltage.

VOLTAGE CONTROLLED OSCILLATOR



V_CC 5.0 V − 14 V
4
8
10 k R1
5
VCO Input
(Input Range 1/3 V_CC to V_CC)
MC1422
3
Output
7
1
2
6
10 k R2
C1 0.01 μF

Circuit Notes

The VCO circuit, which has a nonlinear transfer characteristic, will operate satisfactorily up to 200 kHz. The VCO input range is effective from ⅓ Vcc to Vcc − 2 V, with the highest control voltage producing the lowest output frequency.

Fig. 94-4

SIMPLE VOLTAGE CONTROLLED OSCILLATOR

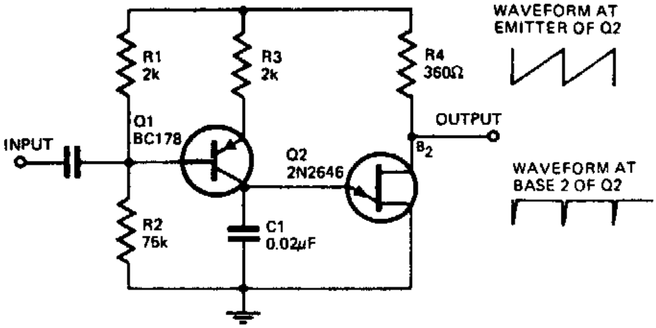

R1
2k
R3
2k
R4
360Ω
WAVEFORM AT
EMITTER OF Q2
OUTPUT
INPUT
Q1
BC178
Q2
2N2646
B2
WAVEFORM AT
BASE 2 OF Q2
R2
75k
C1
0.02µF

Fig. 94-5

Circuit Notes

With the component values shown, the oscillator has a frequency of 8 kHz. When an input signal is applied to the base of Q1 the current flowing through Q1 is varied, thus varying the time required to charge C1. Due to the phase inversion in Q1 the direction of output frequency change is 180 degrees out of phase with the input signal. The output may be used to trigger a bistable flip-flop.

THREE DECADES VCO

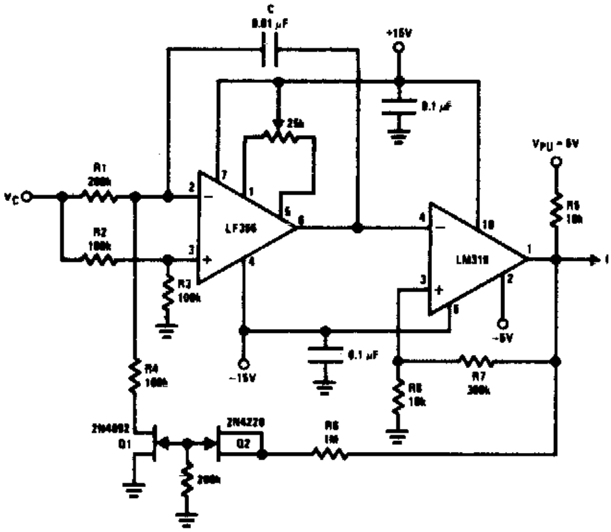
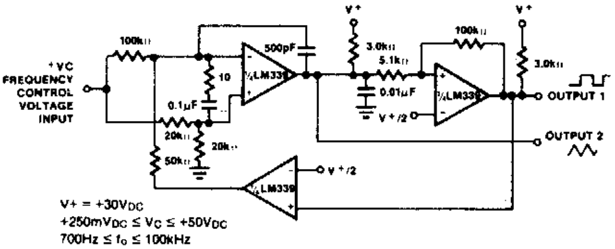

C
0.01 µF
+15V
0.1 µF
25k
V_{PU} = 5V
R1
200k
R2
100k
R3
100k
R4
100k
V_C
LF356
LM311
R5
10k
f
−15V
0.1 µF
−5V
R7
300k
R8
10k
R6
1M
2N4092
Q1
2N4220
Q2
200k
2
3
7
1
5
6
4
4
3
8
10
1
2

Fig. 94-6

$$f = \frac{V_C\,(R8+R7)}{[8\,V_{PU}\,R8\,R1]\,C}, \quad 0 \le V_C \le 30V,\ 10\text{ Hz} \le f \le 10\text{ kHz}$$

R1, R4 matched. Linearity 0.1% over 2 decades.

TWO-DECADE HIGH-FREQUENCY VCO


+VC
FREQUENCY
CONTROL
VOLTAGE
INPUT
100kΩ
10
0.1μF
20kΩ
20kΩ
50kΩ
500pF
¼LM339
V+
3.0kΩ
5.1kΩ
0.01μF
V+/2
100kΩ
¼LM339
V+
3.0kΩ
OUTPUT 1
OUTPUT 2
¼LM339
V+/2

V+ = +30V_{DC}
+250mV_{DC} ≤ V_C ≤ +50V_{DC}
700Hz ≤ f_o ≤ 100kHz

Fig. 94-7

VOLTAGE CONTROLLED OSCILLATOR

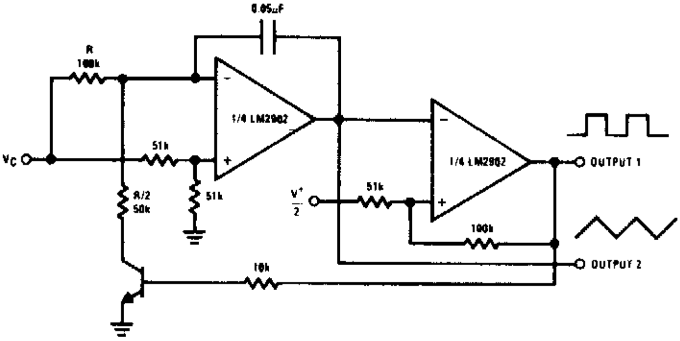
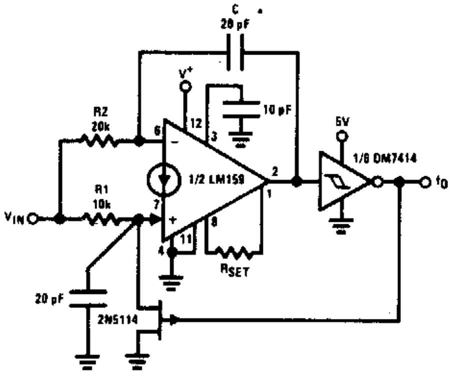

0.05μF
R
100k
1/4 LM2902
V_C
51k
51k
R/2
50k
1/4 LM2902
OUTPUT 1
$\frac{V^+}{2}$
51k
100k
OUTPUT 2
10k

Fig. 94-8

VOLTAGE CONTROLLED OSCILLATOR


C
20 pF
V^+
12
3
10 pF
R2
20k
6
1/2 LM159
2
1
R1
10k
7
V_{IN}
4
11
8
R_{SET}
5V
1/6 DM7414
f_0
20 pF
2N5114

$$f_0 = \frac{V_{IN} - \phi}{4C\Delta V\ R1}$$

where: R2 = 2R1
ϕ = amplifier input voltage ≈ 0.6V
ΔV = DM7414 hysteresis, typ 1V

- 5 MHz operation
- T^2L ouput

Fig. 94-9

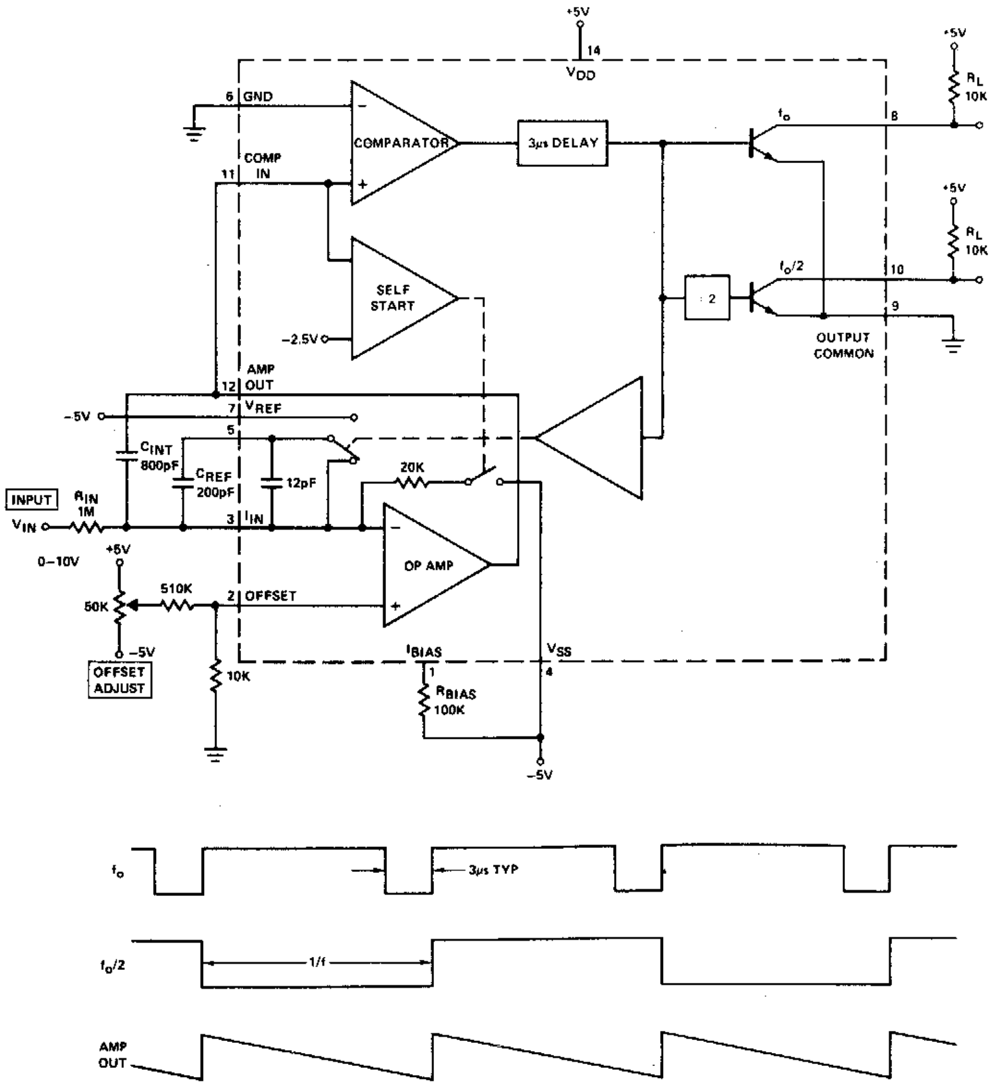
95

Voltage-to-Frequency Converters

The sources of the following circuits are contained in the Sources section beginning on page 730. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

10 Hz to 10 kHz Voltage/Frequency Converter
Voltage-to-Frequency Converter
Voltage-to-Frequency Converter
V/F Conversion, Positive Input Voltage
Ultraprecision V/F Converter
V/F Conversion, Negative Input Voltage

10 Hz TO 10 kHz VOLTAGE/FREQUENCY CONVERTER



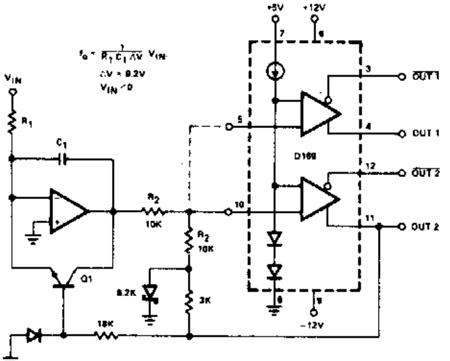
+5V
14
VDD
6 GND
COMPARATOR
3μs DELAY
11 COMP IN
SELF START
−2.5V
f_o
8
+5V
RL 10K
+5V
RL 10K
$f_o/2$
10
÷2
9
OUTPUT COMMON
12 AMP OUT
7 VREF
−5V
5
CINT 800pF
CREF 200pF
12pF
20K
INPUT
RIN 1M
VIN
3 IIN
0–10V
OP AMP
+5V
510K
50K
2 OFFSET
−5V
OFFSET ADJUST
10K
IBIAS
1
RBIAS 100K
VSS
4
−5V
f_o
3μs TYP
$f_o/2$
1/f
AMP OUT

1. To adjust f_{min}, set V_{IN} = 10mV and adjust the 50K offset for 10Hz out.
2. To adjust f_{max}, set V_{IN} 10V and adjust R_{IN} or V_{REF} for 10KHz out.
3. To increase f_{OUT}MAX to 100KHz change C_{REF} to 15pF and C_{INT} to 75pF
4. For high performance applications use high stability components for R_{IN}, C_{REF}, V_{REF} (metal film resistors and glass film capacitors). Also separate the output ground (Pin 9) from the input ground (Pin 6).

Output Waveforms

Fig. 95-1

VOLTAGE-TO-FREQUENCY CONVERTER



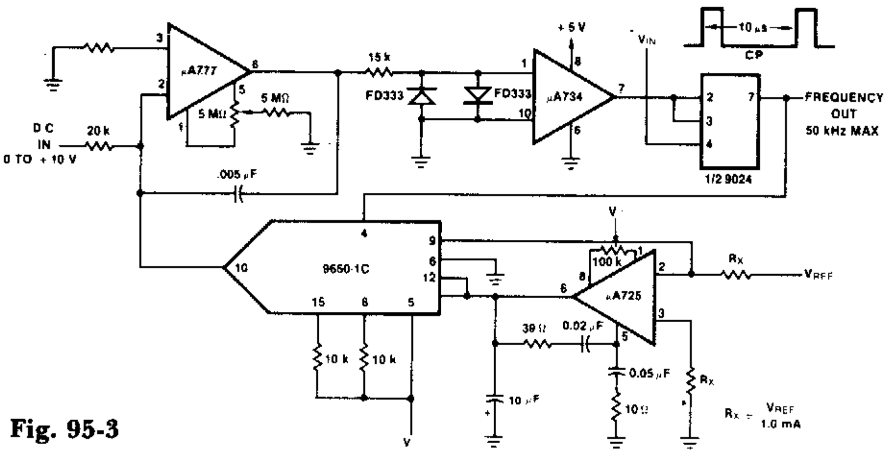
$f_o = \frac{1}{R_1 C_1 \Delta V} V_{IN}$
ΔV = 8.2V
$V_{IN} < 0$
V_{IN}
R_1
C_1
R_2
10K
R_2
10K
Q1
8.2K
3K
10K
+5V
+12V
7
8
3
$\overline{OUT 1}$
5
4
OUT 1
D169
12
$\overline{OUT 2}$
10
11
OUT 2
6
9
−12V

Fig. 95-2

Circuit Notes

The D169 serves as a level detector and provides complementary outputs. The op amp is used to integrate the input signal V_{IN} with a time constant of R1C1. The input (must be negative) causes a positive ramp at the output of the integrator which is summed with a negative zener voltage. When the ramp is positive enough D169 outputs change state and OUT 2 flips from negative to positive. The output pulse repetition rate f_o, is directly proportioned to the negative input voltage V_{IN}.

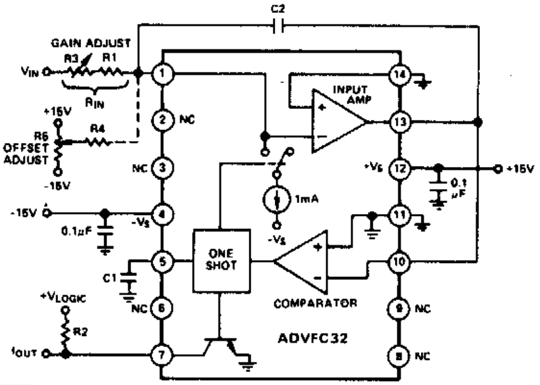
VOLTAGE-TO-FREQUENCY CONVERTER



3
2
μA777
6
5
1
5 MΩ
5 MΩ
DC
IN
0 TO +10 V
20 k
.005 μF
15 k
FD333
FD333
+5 V
1
8
10
μA734
6
7
$-V_{IN}$
10 μs
CP
2
3
4
7
1/2 9024
FREQUENCY
OUT
50 kHz MAX
4
9650-1C
10
9
6
12
15
8
5
10 k
10 k
V
V
100 k
1
2
8
6
μA725
3
5
R_X
V_{REF}
39 Ω
0.02 μF
0.05 μF
10 μF
10 Ω
R_X
$R_X = \frac{V_{REF}}{1.0 \text{ mA}}$

Fig. 95-3

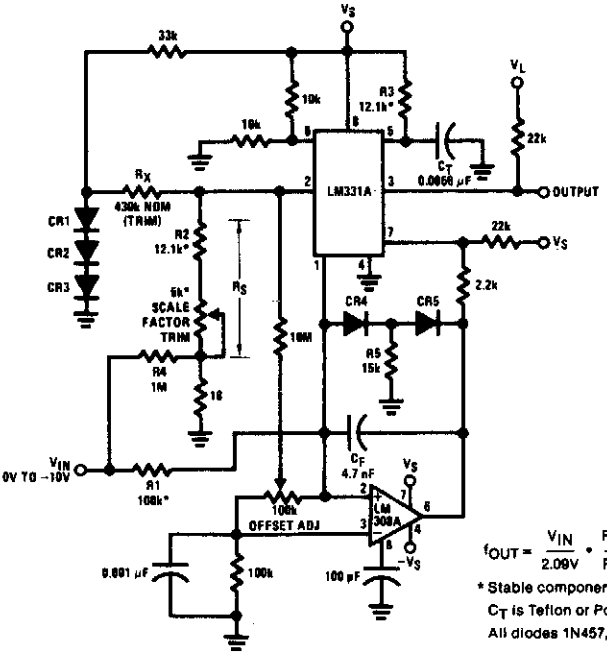
V/F CONVERSION, POSITIVE INPUT VOLTAGE



C2
GAIN ADJUST
R3
R1
V_{IN}
R_{IN}
+15V
R5
OFFSET
ADJUST
R4
−15V
−15V
0.1μF
$-V_S$
C1
$+V_{LOGIC}$
R2
f_{OUT}
1
2
NC
NC
3
4
5
NC
6
7
14
INPUT
AMP
13
$+V_S$
12
+15V
0.1 μF
1mA
$-V_S$
11
10
ONE
SHOT
COMPARATOR
9
NC
8
NC
ADVFC32

Fig. 95-4

ULTRAPRECISION V/F CONVERTER


VS
33k
10k
10k
R3
12.1k*
8
6
5
CT
0.0068 μF
VL
22k
RX
430k NOM
(TRIM)
2
LM331A
3
OUTPUT
CR1
CR2
CR3
R2
12.1k*
RS
7
22k
VS
1
4
2.2k
5k*
SCALE
FACTOR
TRIM
10M
CR4
CR5
R5
15k
R4
1M
10
CF
4.7 nF
VIN
0V TO −10V
R1
100k*
100k
OFFSET ADJ
VS
2
7
LM
308A
6
3
4
8
−VS
0.001 μF
100k
100 pF

$$f_{OUT} = \frac{V_{IN}}{2.09V} \cdot \frac{R_S}{R1} \cdot \frac{1}{R_T C_T}$$

Full-scale output 10 kHz

* Stable components with low tempco; see text

C_T is Teflon or Polystyrene

All diodes 1N457, 1N484, or FD333 (low-leakage silicon)

Fig. 95-5

Circuit Notes

The circuit is capable of better than 0.02% error and 0.003% nonlinearity for a ±20°C range about room temperature.

V/F CONVERSION, NEGATIVE INPUT VOLTAGE

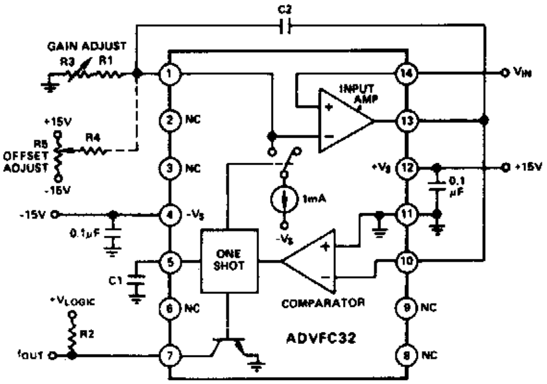

C2
GAIN ADJUST
R3
R1
1
14
VIN
INPUT
AMP
+15V
2
NC
13
R5
OFFSET
ADJUST
R4
3
NC
+VS
12
+15V
-15V
0.1 μF
1mA
-15V
4
-VS
11
0.1μF
-VS
5
ONE
SHOT
10
C1
+VLOGIC
6
NC
COMPARATOR
9
NC
R2
ADVFC32
fOUT
7
8
NC

Fig. 95-6

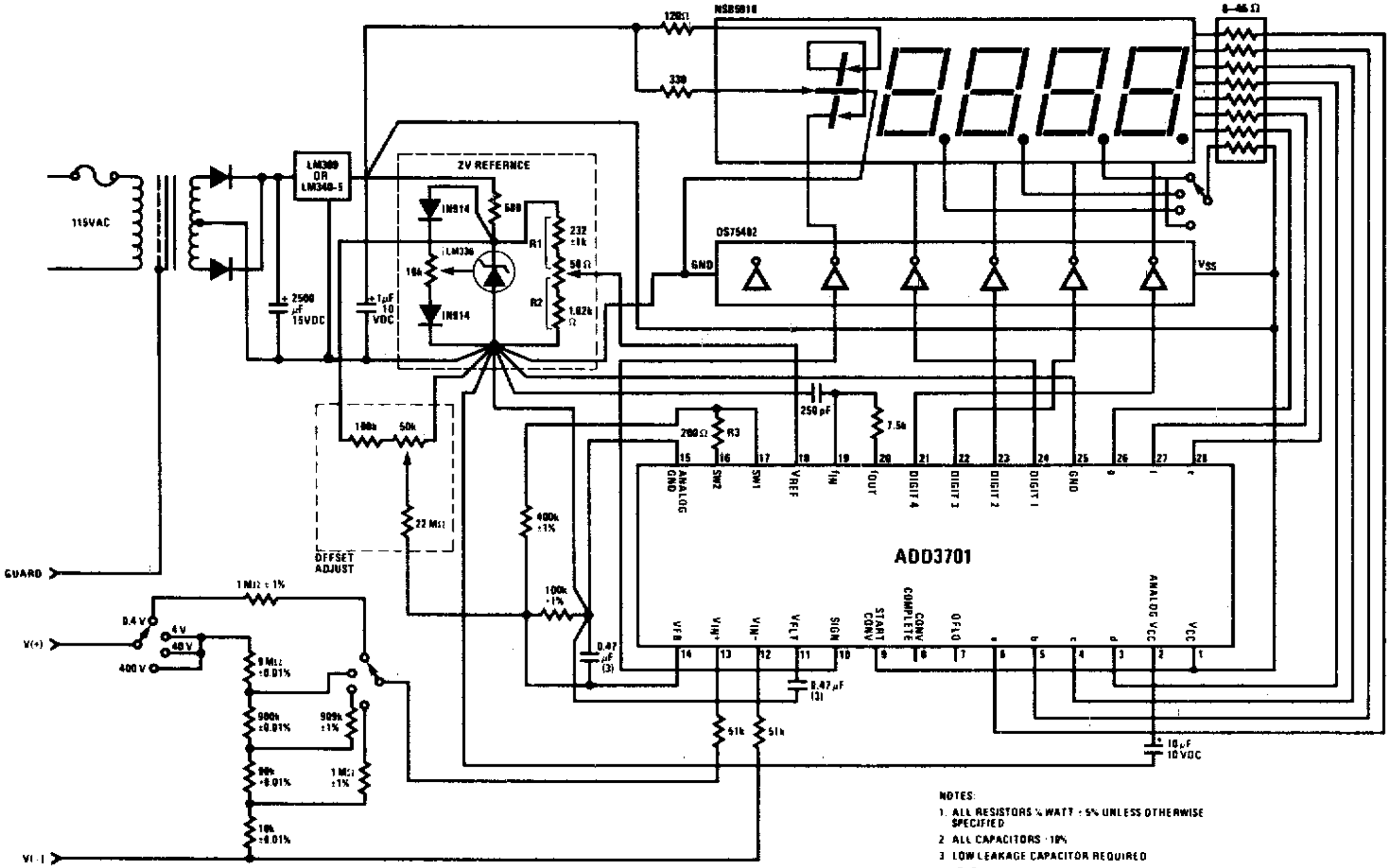
96

Voltmeters

The sources of the following circuits are contained in the Sources section beginning on page 730. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

3-¾ Digit DVM, Four Decade, ±0.4 V, ±4 V, ±40 V, and ±400 V Full Scale
Automatic Nulling DVM
3-½ Digit True RMS AC Voltmeter
3-½ Digit DVM Common Anode Display
DVM Auto-Calibrate Circuit
FET Voltmeter
Extended Range VU Meter (Bar Mode)
High Input Impedance Millivoltmeter
Wide Band AC Voltmeter
Suppressed Zero Meter
Ac Millivoltmeter
4½ Digit LCD-DVM
Sensitive Low Cost VTVM

3-¾ DIGIT DVM, FOUR DECADE, ±0.4 V, ±4 V, ±40 V, AND ±400 V FULL SCALE

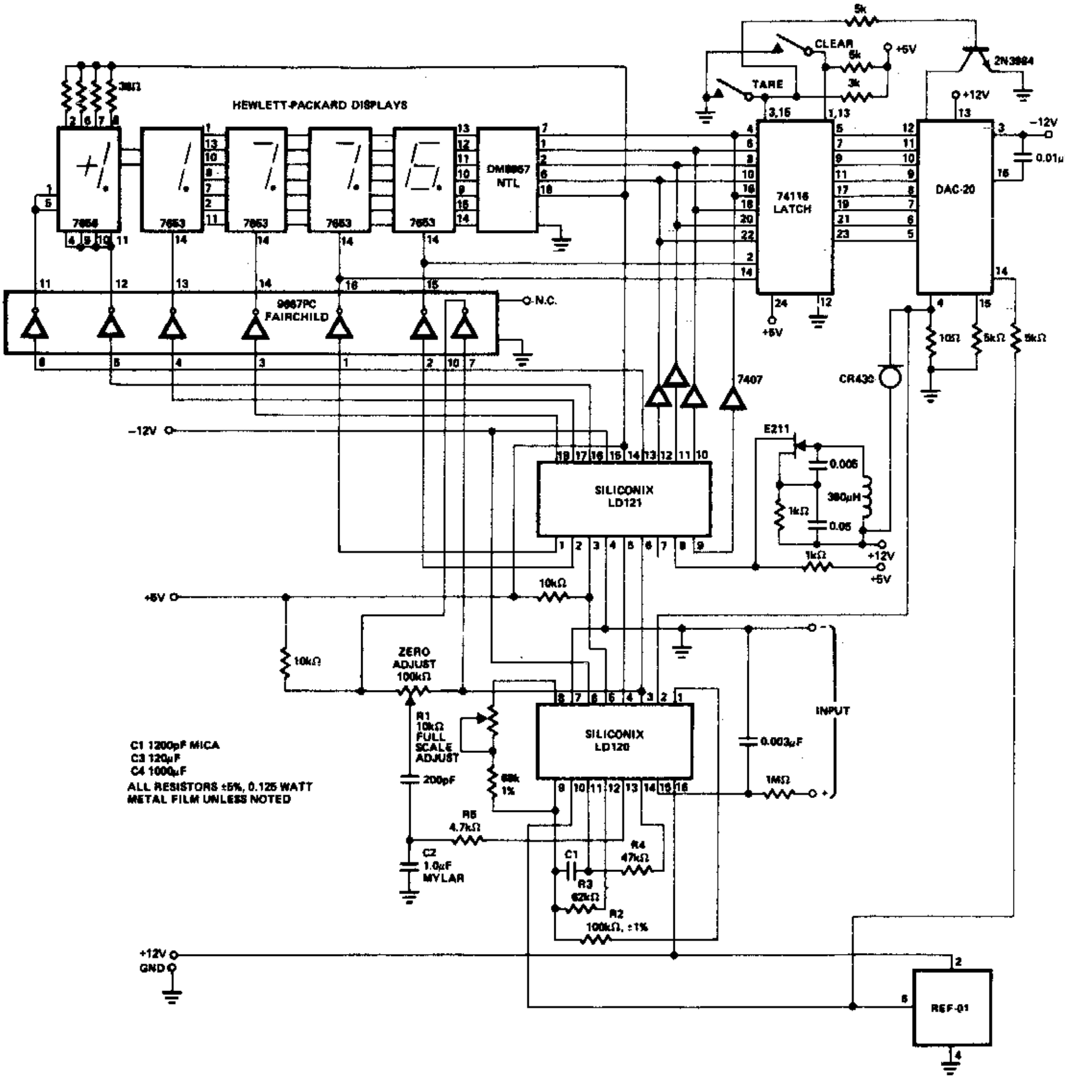

NSB5918
8–46 Ω
120 Ω
330
DS75492
GND
VSS
115VAC
LM309 OR LM340-5
+ 2500 µF 15VDC
+ 1µF 10 VDC
2V REFERENCE
IN914
IN914
580
LM336
10k
R1
232 ±1k
50 Ω
R2
1.82k Ω
100k
50k
22 MΩ
OFFSET ADJUST
250 pF
7.5k
200 Ω
R3
400k ±1%
100k ±1%
0.47 µF (3)
0.47 µF (3)
51k
51k
10 µF 10 VDC
ADD3701
15 ANALOG GND
16 SW2
17 SW1
18 VREF
19 fIN
20 fOUT
21 DIGIT 4
22 DIGIT 3
23 DIGIT 2
24 DIGIT 1
25 GND
26 g
27 f
28 e
14 VFB
13 VIN+
12 VIN-
11 VFLT
10 SIGN
9 START CONV
8 CONV COMPLETE
7 OFLO
6 a
5 b
4 c
3 d
2 ANALOG VCC
1 VCC
GUARD
V(+)
V(-)
0.4 V
4 V
40 V
400 V
1 MΩ ±1%
9 MΩ ±0.01%
900k ±0.01%
90k ±0.01%
10k ±0.01%
909k ±1%
1 MΩ ±1%

NOTES:

1. ALL RESISTORS ¼ WATT ±5% UNLESS OTHERWISE SPECIFIED
2. ALL CAPACITORS ±10%
3. LOW LEAKAGE CAPACITOR REQUIRED
4. $\frac{R_1 R_2}{R_1 + R_2} + R_3 = 25\Omega$

Fig. 96-1

AUTOMATIC NULLING DVM


HEWLETT-PACKARD DISPLAYS
38Ω
7656
7653
7653
7653
7653
DM8857
NTL
9667PC
FAIRCHILD
N.C.
7407
74116
LATCH
DAC-20
CLEAR
TARE
5k
5k
3k
+5V
2N3904
+12V
−12V
0.01µF
10Ω
5kΩ
5kΩ
CR430
E211
0.005
380µH
1kΩ
0.05
1kΩ
+12V
+5V
−12V
SILICONIX
LD121
+5V
10kΩ
10kΩ
ZERO
ADJUST
100kΩ
R1
10kΩ
FULL
SCALE
ADJUST
200pF
68k
1%
SILICONIX
LD120
0.003µF
1MΩ
INPUT
R5
4.7kΩ
C2
1.0µF
MYLAR
C1
R3
62kΩ
R4
47kΩ
R2
100kΩ, ±1%
+12V
GND
REF-01

C1 1200pF MICA
C3 120µF
C4 1000µF
ALL RESISTORS ±5%, 0.125 WATT
METAL FILM UNLESS NOTED

Fig. 96-2

3-½ DIGIT TRUE RMS AC VOLTMETER

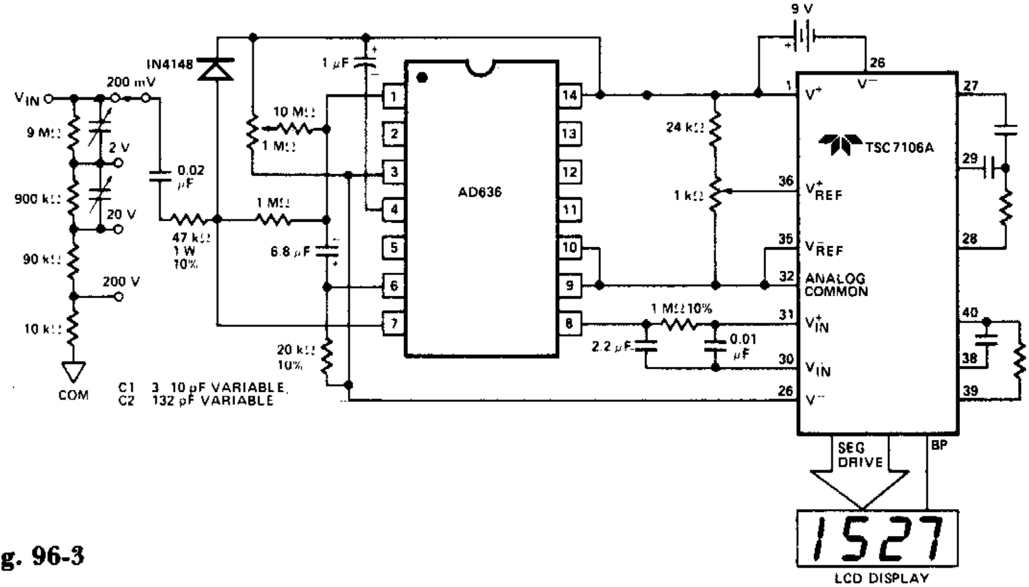

IN4148
1 µF
9 V
V_{IN}
200 mV
9 MΩ
2 V
10 MΩ
1 MΩ
0.02 µF
900 kΩ
20 V
47 kΩ
1 W
10%
1 MΩ
6.8 µF
90 kΩ
200 V
10 kΩ
COM
C1 3 10 pF VARIABLE,
C2 132 pF VARIABLE
20 kΩ
10%
AD636
24 kΩ
1 kΩ
TSC7106A
V^+
V^-
V^+_{REF}
V^-_{REF}
ANALOG COMMON
1 MΩ 10%
2.2 µF
0.01 µF
V^+_{IN}
V^-_{IN}
V^-
SEG DRIVE
BP
1527
LCD DISPLAY

Fig. 96-3

3½ DIGIT DVM (±200.0 mV) COMMON ANODE DISPLAY

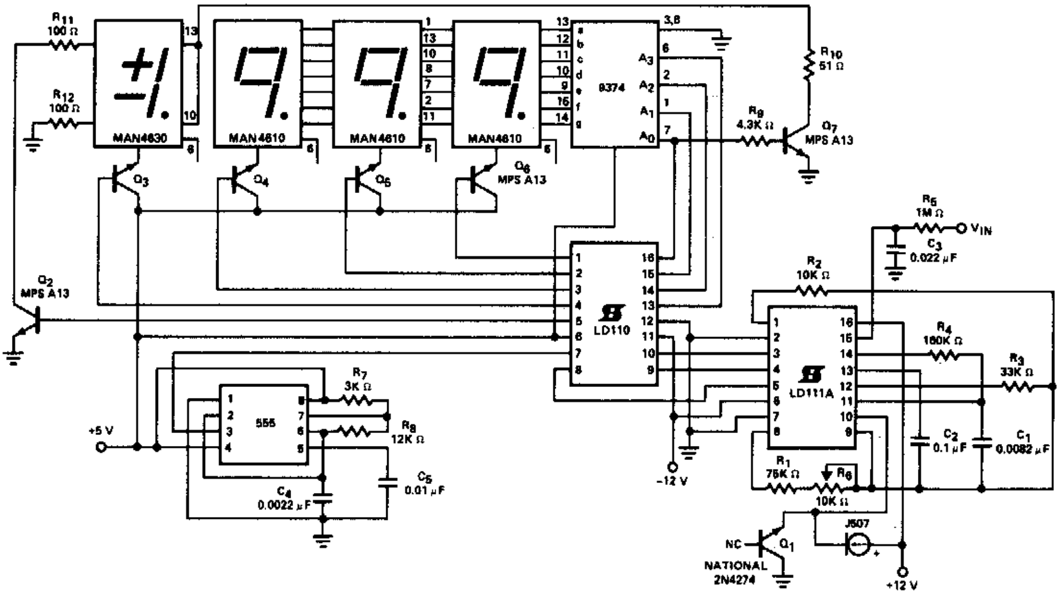

R_{11} 100 Ω
R_{12} 100 Ω
MAN4630
MAN4610
MAN4610
MAN4610
8374
A_3
A_2
A_1
A_0
3,8
R_{10} 51 Ω
R_9 4.3K Ω
Q_7 MPS A13
Q_3
Q_4
Q_5
Q_6 MPS A13
Q_2 MPS A13
LD110
R_5 1M Ω
V_{IN}
C_3 0.022 µF
R_2 10K Ω
LD111A
R_4 180K Ω
R_3 33K Ω
555
R_7 3K Ω
R_8 12K Ω
+5 V
C_4 0.0022 µF
C_5 0.01 µF
−12 V
C_2 0.1 µF
C_1 0.0082 µF
R_1 75K Ω
R_6 10K Ω
J507
NC
Q_1
NATIONAL 2N4274
+12 V

Fig. 96-4

DVM AUTO-CALIBRATE CIRCUIT

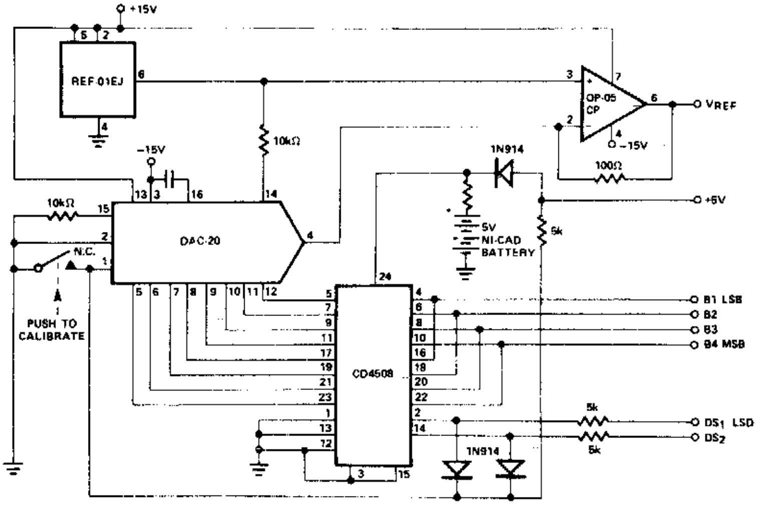

+15V
REF-01EJ
5
2
6
4
3
7
OP-05 CP
6
VREF
2
4
−15V
100Ω
10kΩ
−15V
1N914
+6V
13
3
16
14
10kΩ
15
DAC-20
2
4
N.C.
1
PUSH TO CALIBRATE
5V NI-CAD BATTERY
5k
24
5
6
7
8
9
10
11
12
7
9
11
17
19
21
23
1
13
12
CD4508
4
6
8
10
16
18
20
22
2
14
3
15
B1 LSB
B2
B3
B4 MSB
5k
DS1 LSD
DS2
5k
1N914

Fig. 96-5

FET VOLTMETER

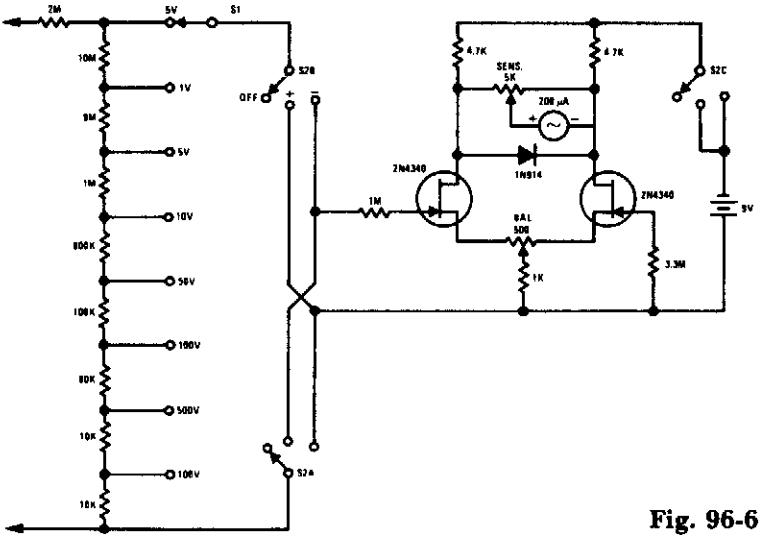
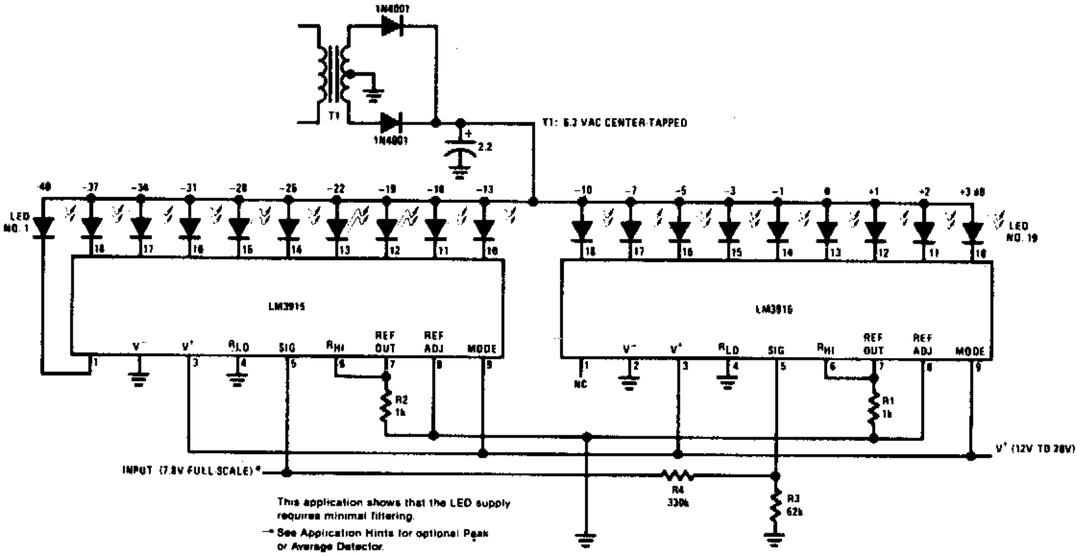

2M
5V
S1
10M
1V
9M
5V
1M
10V
900K
50V
100K
100V
90K
500V
10K
1000V
10K
S2B
OFF
+
−
S2A
1M
2N4340
4.7K
4.7K
SENS. 5K
200 µA
+
−
1N914
2N4340
BAL 500
1K
3.3M
S2C
9V

Fig. 96-6

Circuit Notes

This FETVM replaces the function of the VTVM while at the same time ridding the instrument of the usual line cord. In addition, drift rates are far superior to vacuum tube circuits allowing a 0.5 volt full scale range which is impractical with most vacuum tubes. The low-leakage, low-noise 2N4340 is an ideal device for this application.

EXTENDED RANGE VU METER (BAR MODE)


1N4001
T1
1N4001
2.2
T1: 6.3 VAC CENTER-TAPPED
-40
-37
-34
-31
-28
-25
-22
-19
-16
-13
-10
-7
-5
-3
-1
0
+1
+2
+3 dB
LED NO. 1
LED NO. 19
LM3915
LM3916
V⁻
V⁺
RLO
SIG
RHI
REF OUT
REF ADJ
MODE
NC
R2 1k
R1 1k
V⁺ (12V TO 28V)
INPUT (7.8V FULL SCALE)*
R4 330k
R3 62k

This application shows that the LED supply requires minimal filtering.

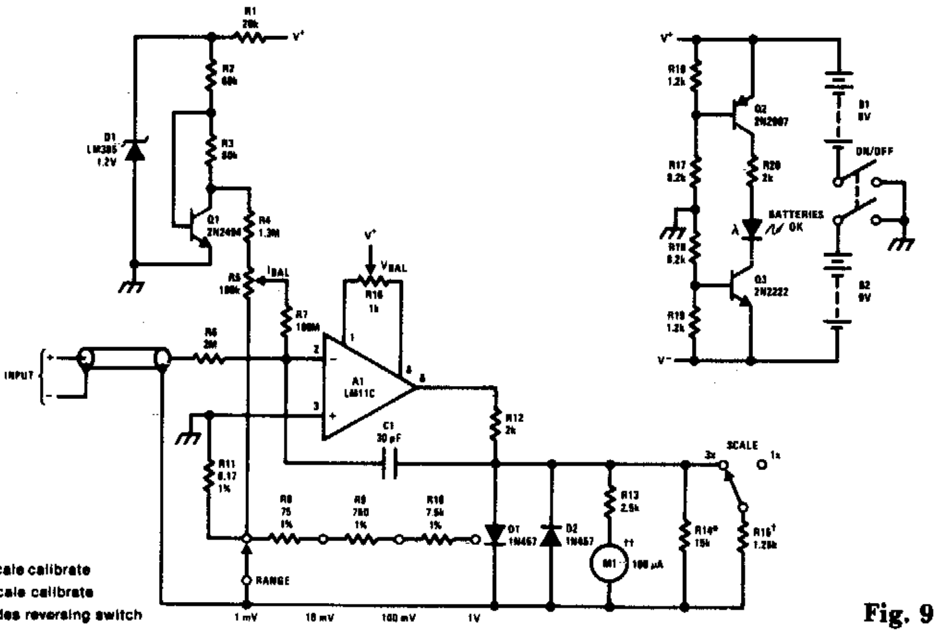
* See Application Hints for optional Peak or Average Detector.

† Adjust R3 for 3 dB difference between LED #11 and LED #12

$$\frac{R3}{R3 + R4} = 0.158 = -16\ dB$$

Fig. 96-7

HIGH INPUT IMPEDANCE MILLIVOLTMETER


R1 20k
V⁺
R2 68k
D1 LM385 1.2V
R3 68k
Q1 2N2484
R4 1.3M
R5 100k
IBAL
V⁺
VBAL
R16 1k
R7 100M
INPUT
R6 2M
A1 LM11C
C1 30 pF
R12 2k
R11 8.17 1%
R8 75 1%
R9 750 1%
R10 7.5k 1%
D1 1N457
D2 1N457
R13 2.5k
M1 100 μA
††
R14* 15k
R15† 1.25k
SCALE
3x
1x
RANGE
1 mV
10 mV
100 mV
1V
V⁺
R18 1.2k
Q2 2N2907
B1 9V
ON/OFF
R17 8.2k
R20 2k
BATTERIES OK
R18 8.2k
Q3 2N2222
B2 9V
R19 1.2k
V⁻

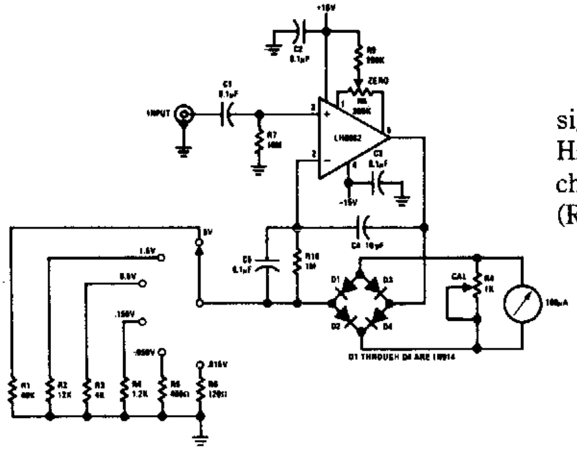
* 1× scale calibrate

† 3× scale calibrate

‡ Includes reversing switch

Fig. 96-8

WIDE BAND AC VOLTMETER

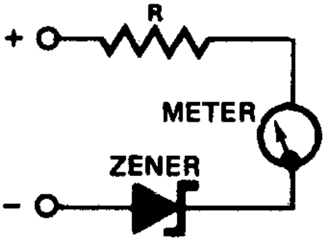

+15V
C2
0.1µF
ZERO
C1
0.1µF
INPUT
R7
LH0062
C3
0.1µF
-15V
5V
1.5V
5.0V
.150V
.050V
.015V
C5
0.1µF
R10
1M
C4 10pF
D1
D2
D3
D4
CAL
R8
1K
100µA
D1 THROUGH D4 ARE 1N914
R1
R2
12K
R3
4K
R4
1.2K
R5
400Ω
R6
120Ω

Circuit Notes

This voltmeter is capable of measuring ac signals as low as 15 mV at frequencies from 100 Hz to 500 kHz. Full scale sensitivity may be changed by altering the values R1 through R6 ($R = V_{IN}/100\ \mu A$).

Fig. 96-9

SUPPRESSED ZERO METER


R
+
METER
ZENER
-

Circuit Notes

A zener diode placed in series with a voltmeter will prevent the meter from reading until the applied voltage exceeds the zener voltage. Thus, a 10 volt zener in series with a 5-volt meter will allow the condition of a 12 V car battery to be monitored with much greater sensitivity than would be possible with a meter reading 0-15 volts.

Fig. 96-10

AC MILLIVOLTMETER

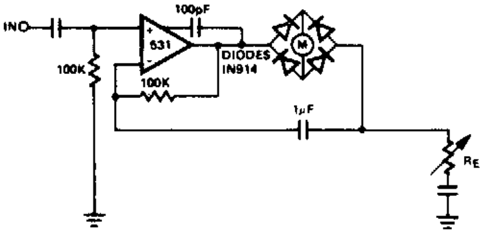

IN
100pF
531
+
-
100K
100K
DIODES
IN914
M
1µF
RE

Fig. 96-11

4½ DIGIT LCD-DVM

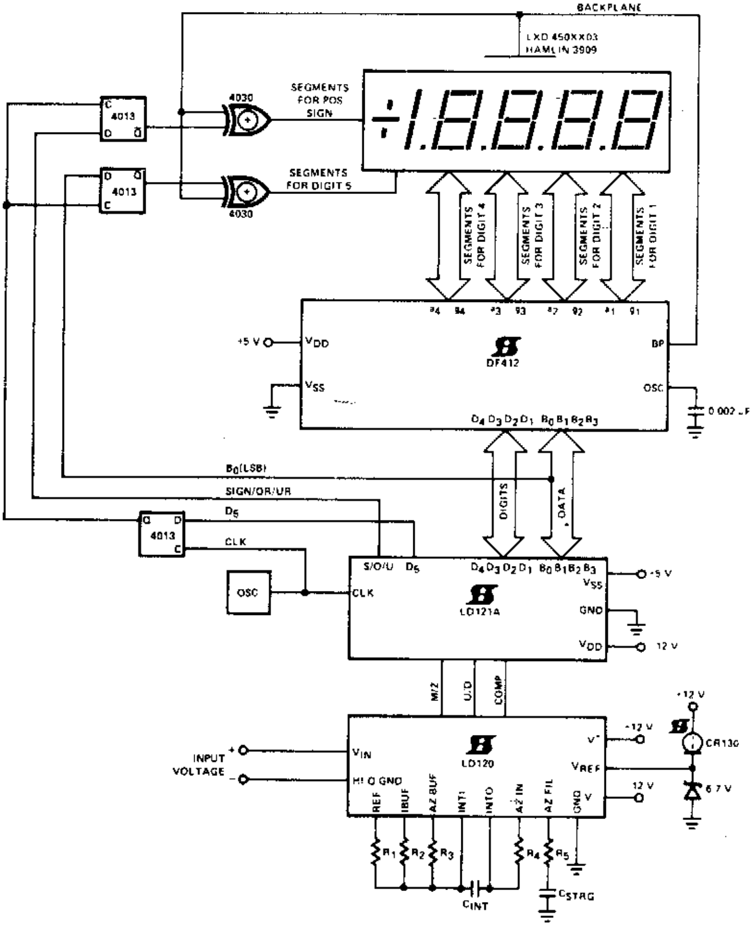

BACKPLANE
LXD 450XX03
HAMLIN 3909
4030
SEGMENTS FOR POS SIGN
4013
C
D
Q
Q̄
SEGMENTS FOR DIGIT 5
SEGMENTS FOR DIGIT 4
SEGMENTS FOR DIGIT 3
SEGMENTS FOR DIGIT 2
SEGMENTS FOR DIGIT 1
a4 b4 a3 b3 a2 b2 a1 b1
+5 V
VDD
VSS
DF412
BP
OSC
0.002 µF
D4 D3 D2 D1 B0 B1 B2 B3
B0(LSB)
SIGN/OR/UR
D5
CLK
DIGITS
DATA
OSC
S/O/U
LD121A
+5 V
VSS
GND
VDD
12 V
M/Z
U/D
COMP
INPUT VOLTAGE
+
−
VIN
HI Q GND
LD120
V+
+12 V
VREF
12 V
V
+12 V
CR130
6.7 V
REF
IBUF
AZ BUF
INT1
INTO
AZ IN
AZ FIL
GND
R1
R2
R3
R4
R5
CINT
CSTRG

Fig. 96-12

SENSITIVE LOW COST "VTVM"

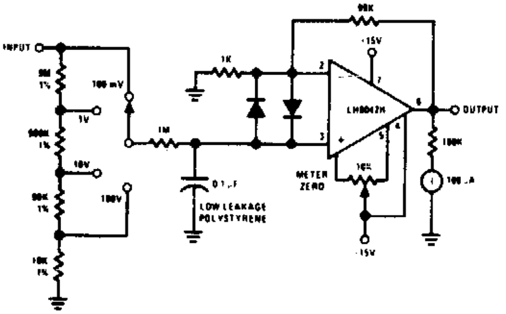

INPUT
9M 1%
100 mV
1V
900K 1%
10V
90K 1%
100V
10K 1%
1M
0.1 µF
LOW LEAKAGE POLYSTYRENE
1K
99K
+15V
2
3
7
6
5
4
LH0042H
METER ZERO
10K
−15V
OUTPUT
100K
100 µA

Fig. 96-13

97

Waveform and Function Generators

The sources of the following circuits are contained in the Sources section beginning on page 730. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

Low Cost Adjustable Function Generator
DAC Controlled Function Generator
Programmed Function Generator
100-kHz Quadrature Oscillator
Strobe-Tone Burst Generator
Low Cost High Frequency Generator
Tone-Burst Oscillator and Decoder
Triangle and Square Waveform Generator
10 kHz Oscillator
50 kHz Oscillator
Variable Audio Oscillator, 20 Hz to 20 kHz
Gated Oscillator
Exponential Digitally-Controlled Oscillator
Function Generator
Clock Source
Precision Oscillator with 20 ns Switching
Oscillator with Quadrature Output
Wide Range Variable Oscillator
Frequency Divider and Staircase Generator
Precision Oscillator to Switch 100 mA Loads

LOW COST ADJUSTABLE FUNCTION GENERATOR

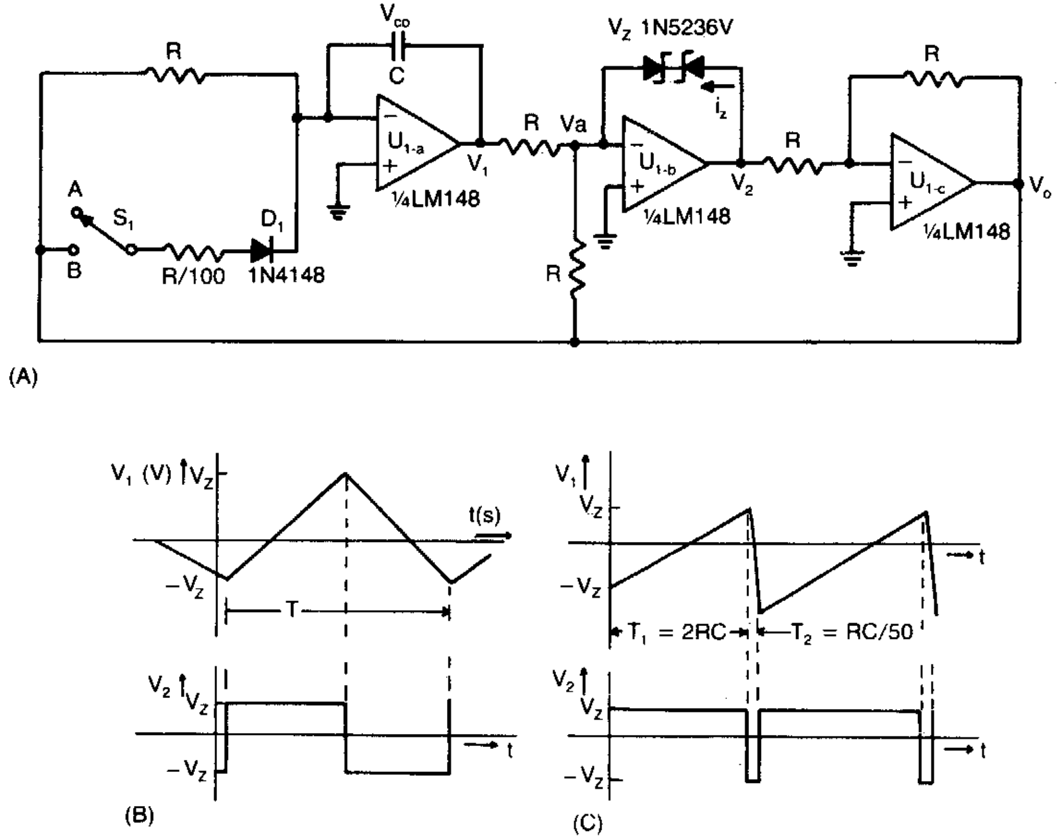
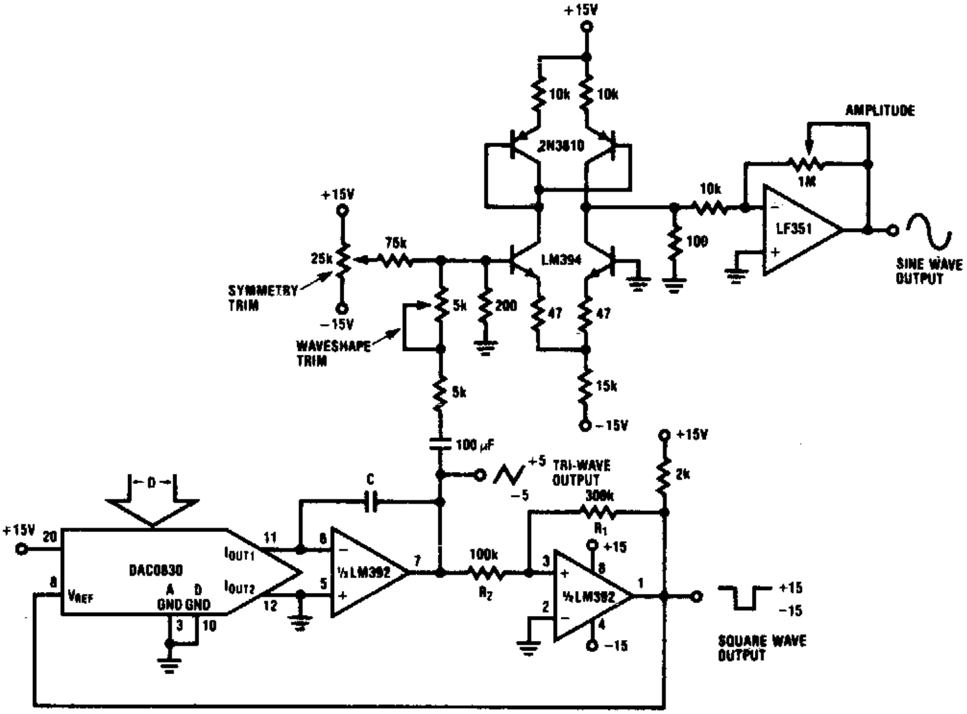

R
V_co
C
V_Z 1N5236V
i_z
R
U_1-a
¼LM148
V_1
R Va
U_1-b
¼LM148
V_2
R
U_1-c
¼LM148
V_o
A
S_1
B
R/100
D_1
1N4148
R
(A)
V_1 (V)
V_Z
-V_Z
t(s)
T
V_2
V_Z
-V_Z
t
(B)
V_1
V_Z
-V_Z
t
T_1 = 2RC
T_2 = RC/50
V_2
V_Z
-V_Z
t
(C)

Fig. 97-1

Circuit Notes

This low-cost operational-amplifier circuit (A) generates four different functions with adjustable periods. For the components shown here, the period of the output waveforms is given by T = 4RC and T = 2RC. With switch S1 in position A, V1 is a triangular waveform, while V2 is a square wave (B). With the switch in position B, a sawtooth waveform is generated at V1 and a pulse at V2(C).

DAC CONTROLLED FUNCTION GENERATOR


+15V
10k
10k
2N3810
AMPLITUDE
1M
10k
LF351
SINE WAVE OUTPUT
100
LM394
47
47
15k
−15V
+15V
25k
75k
SYMMETRY TRIM
−15V
5k
200
WAVESHAPE TRIM
5k
100 μF
+5
−5
TRI-WAVE OUTPUT
+15V
2k
300k
R_1
C
D
+15V
20
DAC0830
V_{REF}
8
A GND
D GND
3
10
I_{OUT1}
I_{OUT2}
11
12
6
5
½ LM392
7
100k
R_2
3
2
8
4
+15
−15
½ LM392
1
+15
−15
SQUARE WAVE OUTPUT

- DAC controls the frequency of sine, square, and triangle outputs.
- $f = \frac{D}{256(20k)C}$ for $V_{OMAX} = V_{OMIN}$ of square wave output and $R_1 = 3R_2$.
- 255 to 1 linear frequency range; oscillator stops with D = 0
- Trim symmetry and wave-shape for minimum sine wave distortion.

Fig. 97-2

PROGRAMMED FUNCTION GENERATOR

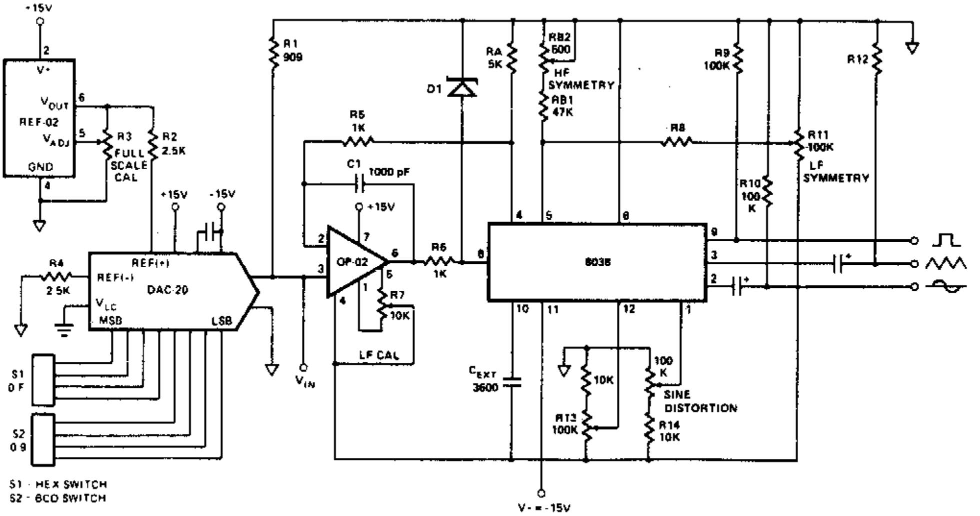
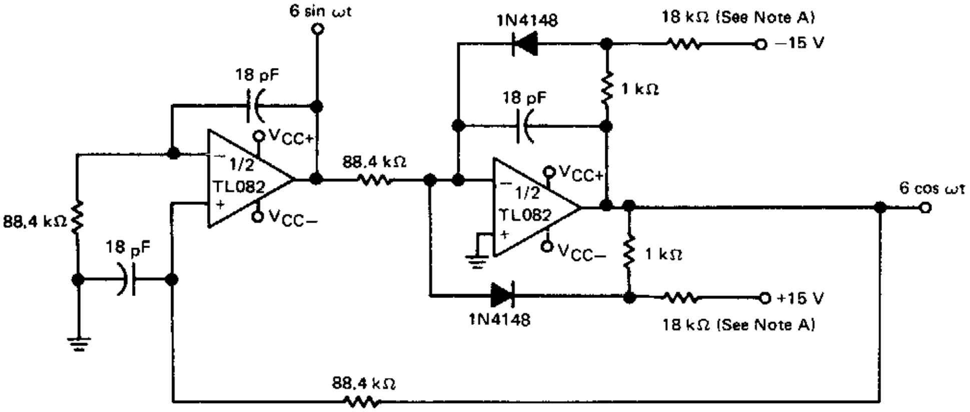

+15V
2
V+
VOUT
6
REF-02
5
VADJ
GND
4
R3
FULL SCALE CAL
R2
2.5K
+15V
-15V
R4
2.5K
REF(+)
REF(-)
DAC-20
VLC
MSB
LSB
S1
0 F
S2
0 9
S1 - HEX SWITCH
S2 - BCD SWITCH
R1
909
VIN
R5
1K
C1
1000 pF
+15V
OP-02
2
3
7
6
4
1
5
R7
10K
LF CAL
R6
1K
D1
RA
5K
RB2
500
HF SYMMETRY
RB1
47K
R8
R9
100K
R10
100
K
R11
100K
LF SYMMETRY
R12
8038
4
5
6
8
9
3
2
10
11
12
1
CEXT
3600
10K
RT3
100K
100
K
SINE DISTORTION
R14
10K
V- = -15V

Fig. 97-3

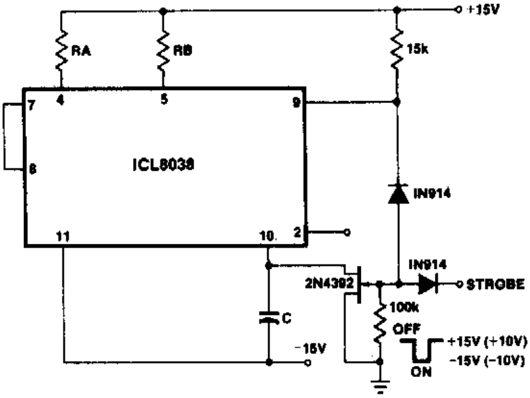
100-kHz QUADRATURE OSCILLATOR


6 sin ωt
18 pF
1/2
TL082
VCC+
VCC−
88.4 kΩ
88.4 kΩ
18 pF
1N4148
18 kΩ (See Note A)
−15 V
1 kΩ
18 pF
1/2
TL082
VCC+
VCC−
6 cos ωt
1 kΩ
1N4148
+15 V
18 kΩ (See Note A)
88.4 kΩ

Note A: These resistor values may be adjusted for a symmetrical output.

Fig. 97-4

STROBE-TONE BURST GENERATOR

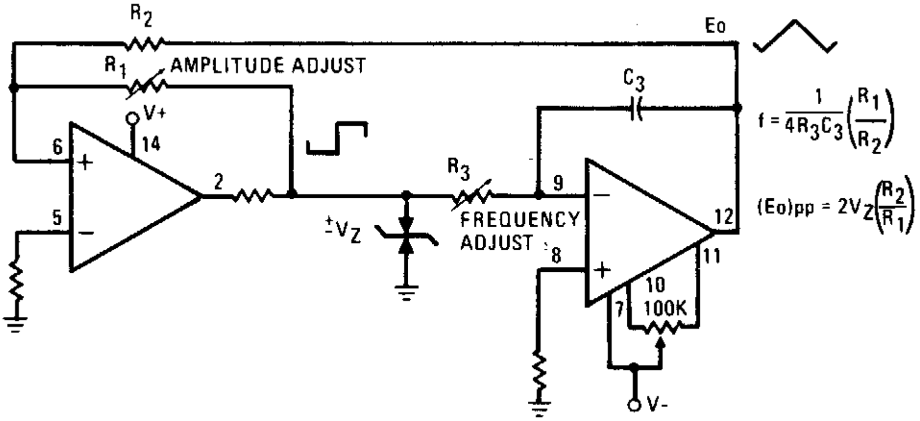

+15V
RA
RB
15k
7
4
5
9
8
ICL8038
IN914
11
10
2
IN914
2N4392
STROBE
100k
OFF
C
-15V
+15V (+10V)
-15V (-10V)
ON

Circuit Notes

With a dual supply voltage, the external capacitor on pin 10 can be shorted to ground to halt the 8038 oscillation. The circuit uses a FET switch and diode ANDed with an input strobe signal to allow the output to always start on the same slope.

Fig. 97-5

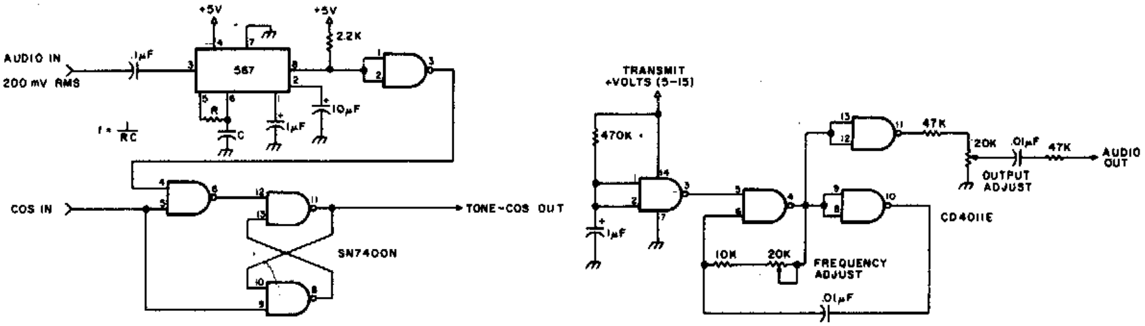
LOW COST HIGH FREQUENCY GENERATOR


R_2
Eo
R_1
AMPLITUDE ADJUST
C_3
V+
14
6
+
5
−
2
R_3
$\pm V_Z$
FREQUENCY ADJUST
9
−
12
8
+
11
10
7
100K
V-
$f = \frac{1}{4R_3C_3}\left(\frac{R_1}{R_2}\right)$
$(E_o)_{pp} = 2V_Z\left(\frac{R_2}{R_1}\right)$

IC =HA2650/26555

Fig. 97-6

TONE-BURST OSCILLATOR AND DECODER


AUDIO IN
200 mV RMS
1μF
+5V
567
+5V
2.2K
R
C
1μF
10μF
f = 1/RC
COS IN
TONE-COS OUT
SN7400N
TRANSMIT
+VOLTS (5-15)
470K
1μF
47K
20K
.01μF
47K
AUDIO OUT
OUTPUT ADJUST
CD4011E
10K
20K
FREQUENCY ADJUST
.01μF

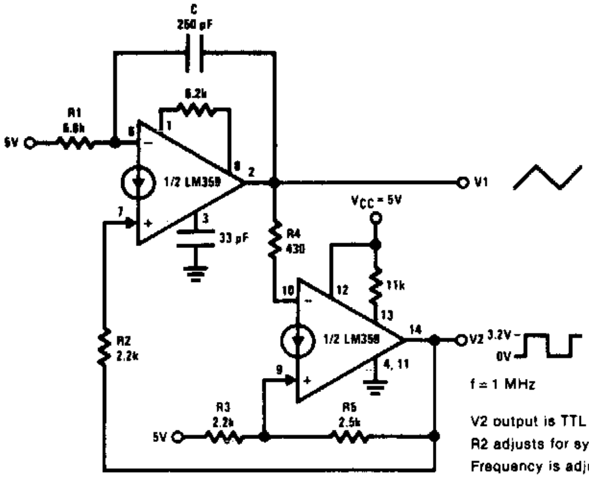
Decoder and logic. *Tone-burst oscillator.*

Fig. 97-7

Circuit Notes

A tone burst sent at the beginning of each transmission is decoded (at receiver) by a PLL causing output from pin 3 of logic gate to turn on carrier-operated switch (COS).

TRIANGLE AND SQUARE WAVEFORM GENERATOR

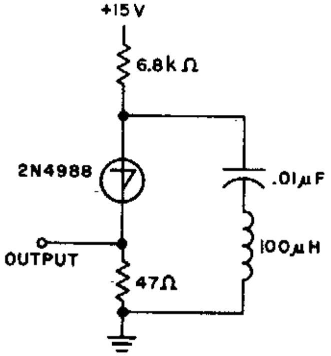

C
250 pF
R1
6.8k
5V
6.2k
1/2 LM358
33 pF
V1
VCC = 5V
R4
430
11k
1/2 LM358
V2
3.2V
0V
R2
2.2k
R3
2.2k
5V
R5
2.5k

f ≈ 1 MHz

V2 output is TTL compatible
R2 adjusts for symmetry of the triangle waveform
Frequency is adjusted with R5 and C

Fig. 97-8

10 kHz OSCILLATOR


+15 V
6.8k Ω
2N4988
.01 μF
100 μH
OUTPUT
47 Ω

Circuit Notes

The capacitor charges until switching voltage is reached. When SUS switches on, the inductor causes current to ring. When the current thru SUS drops below the holding current, the device turns off and the cycle repeats.

Fig. 97-9

50 kHz OSCILLATOR

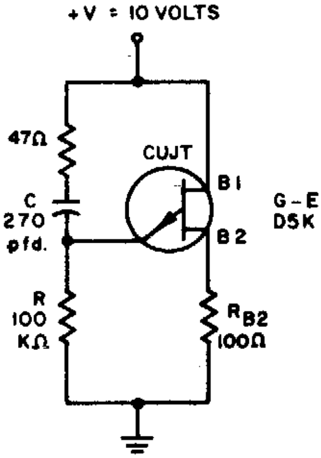

+V = 10 VOLTS
47Ω
CUJT
B1
G-E
D5K
C
270
pfd.
B2
R
100
KΩ
RB2
100Ω

Fig. 97-10

Circuit Notes

A 50 kHz circuit is possible because of the more nearly ideal characteristics of the D5K.

VARIABLE AUDIO OSCILLATOR, 20 Hz TO 20 kHz

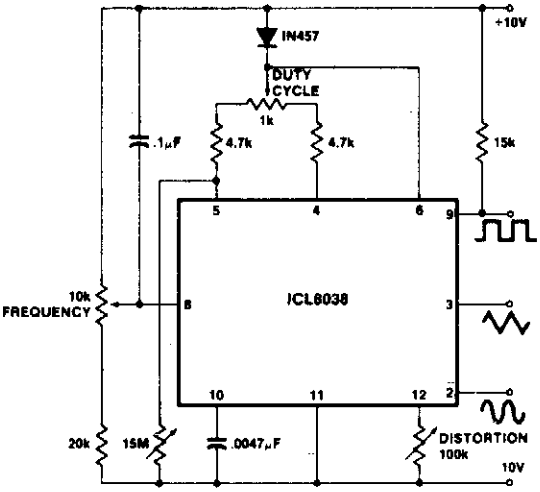

+10V
IN457
DUTY CYCLE
1k
.1μF
4.7k
4.7k
15k
5
4
6
9
10k FREQUENCY
8
ICL8038
3
10
11
12
2
20k
15M
.0047μF
DISTORTION 100k
10V

Fig. 97-11

Circuit Notes

To obtain a 1000:1 Sweep Range, the voltage across external resistors R_A and R_B must decrease to nearly zero. This requires that the highest voltage on control pin 8 exceed the voltage at the top of R_A and R_B by a few hundred millivolts. The circuit achieves this by using a diode to lower the effective supply voltage on the 8038. The large resistor on pin 5 helps reduce duty cycle variations with sweep.

GATED OSCILLATOR

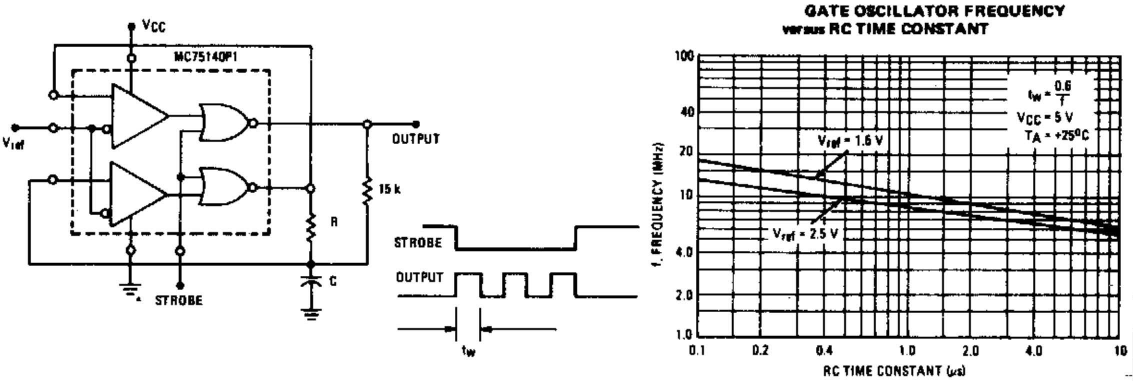

VCC
MC75140P1
Vref
OUTPUT
15 k
R
C
STROBE
STROBE
OUTPUT
tw
GATE OSCILLATOR FREQUENCY versus RC TIME CONSTANT
f, FREQUENCY (MHz)
100
40
20
10
4.0
2.0
1.0
0.1
0.2
0.4
1.0
2.0
4.0
10
RC TIME CONSTANT (μs)
Vref = 1.6 V
Vref = 2.5 V
tw = 0.6/f
VCC = 5 V
TA = +25°C

Fig. 97-12

EXPONENTIAL DIGITALLY-CONTROLLED OSCILLATOR

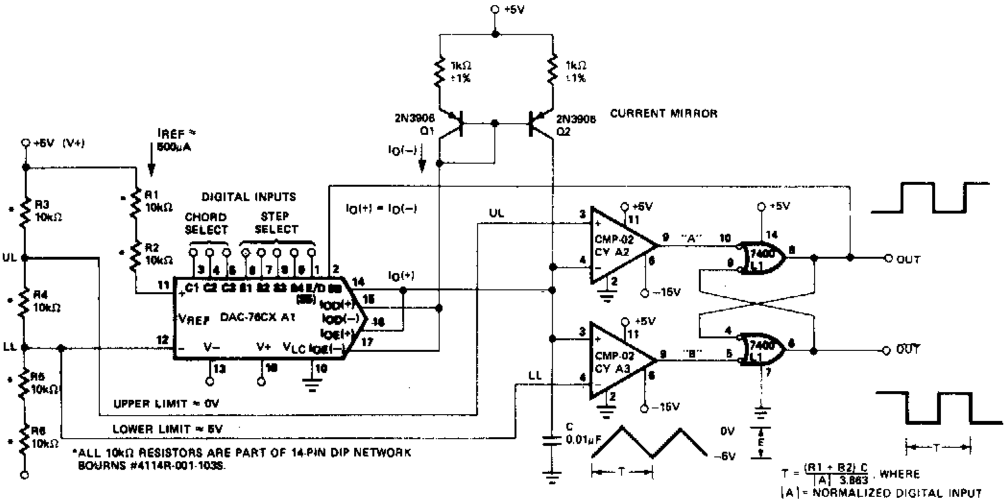
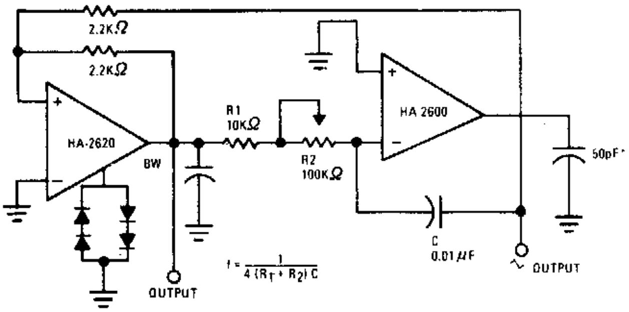

+5V
1kΩ ±1%
1kΩ ±1%
2N3906 Q1
2N3906 Q2
CURRENT MIRROR
IO(−)
+5V (V+)
IREF ≈ 500μA
R3 10kΩ
R1 10kΩ
R2 10kΩ
UL
R4 10kΩ
LL
R5 10kΩ
R6 10kΩ
DIGITAL INPUTS
CHORD SELECT
STEP SELECT
C1 C2 C3 S1 S2 S3 S4 E/D SB (SB)
VREF
DAC-76CX A1
IOD(+)
IOD(−)
IOE(+)
IOE(−)
V−
V+
VLC
IO(+) = IO(−)
IO(+)
UL
LL
CMP-02 CY A2
CMP-02 CY A3
"A"
"B"
+5V
−15V
7400 ¼
OUT
OUT
UPPER LIMIT ≈ 0V
LOWER LIMIT ≈ 5V
C 0.01μF
0V
−5V
T
*ALL 10kΩ RESISTORS ARE PART OF 14-PIN DIP NETWORK BOURNS #4114R-001-103S.
T = (R1 + R2) C / |A| 3.863, WHERE
|A| = NORMALIZED DIGITAL INPUT

Fig. 97-13

Circuit Notes

The microprocessor-controlled oscillator has a 8159 to 1 frequency range covering 2.5 Hz to 20 kHz. An exponential, current output IC DAC functioning as a programmable current source alternately charges and discharges a capacitor between precisely-controlled upper and lower limits. The circuit features instantaneous frequency change, operates with +5 ±1 V and −15 V ±3 V supplies, and has the dynamic range of a 13-bit DAC.

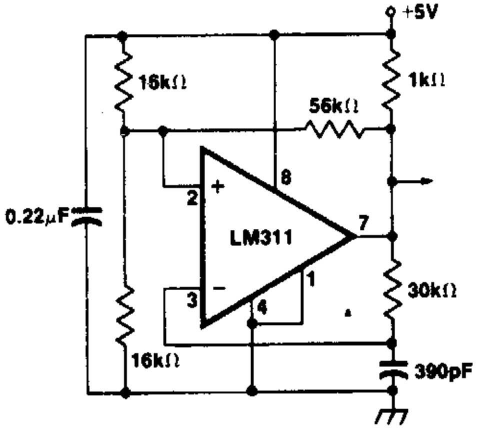
FUNCTION GENERATOR


2.2KΩ
2.2KΩ
HA-2620
BW
R1
10KΩ
R2
100KΩ
HA 2600
50pF
C
0.01μF
OUTPUT
OUTPUT
f = 1 / 4 (R_T + R_2) C

IC = HA-2620

Fig. 97-14

CLOCK SOURCE


+5V
16kΩ
1kΩ
56kΩ
8
2
+
LM311
7
0.22μF
3
−
4
1
30kΩ
16kΩ
390pF

Circuit Notes

A clock source using LM311 voltage comparator in positive feedback mode to minimize clock frequency shift problem.

Fig. 97-15

PRECISION OSCILLATOR WITH 20 NS SWITCHING

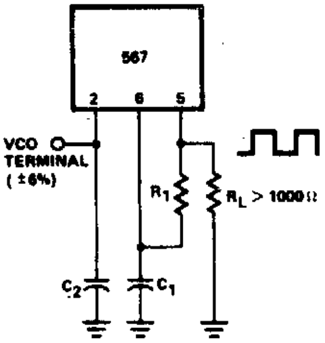

567
2
6
5
VCO TERMINAL (±6%)
R_1
R_L > 1000Ω
C_2
C_1

Fig. 97-16

OSCILLATOR WITH QUADRATURE OUTPUT

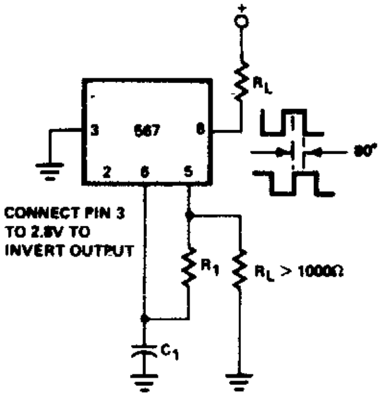
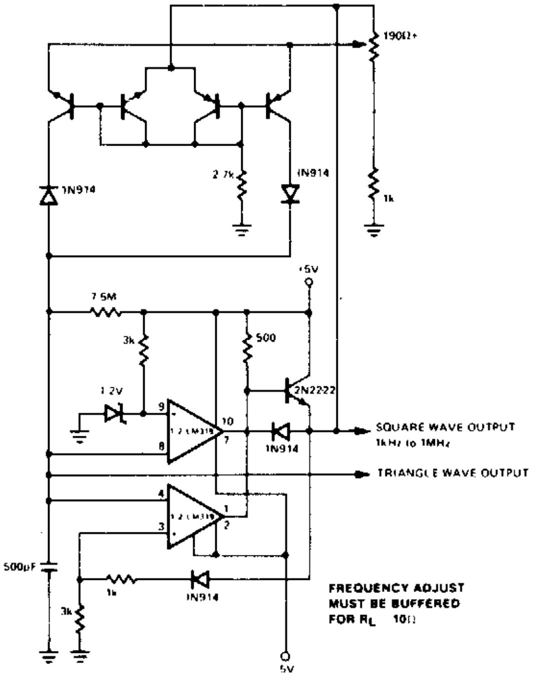

+
R_L
3
567
8
2
6
5
90°
CONNECT PIN 3 TO 2.8V TO INVERT OUTPUT
R_1
R_L > 1000Ω
C_1

Fig. 97-17

WIDE RANGE VARIABLE OSCILLATOR

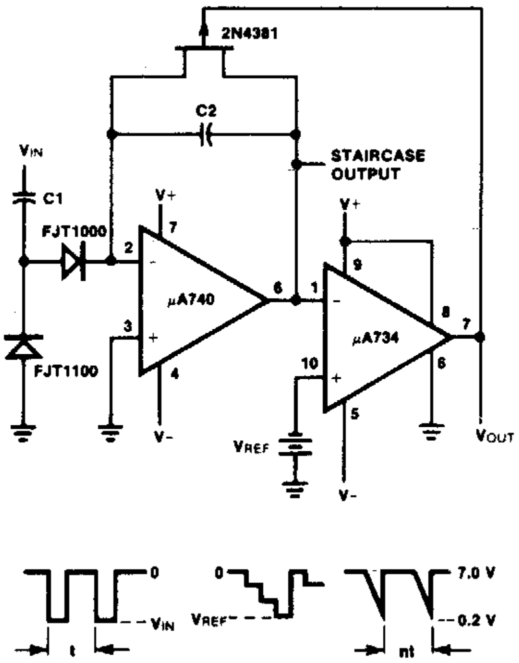


190Ω+
1N914
2.7k
1N914
1k
+5V
7.5M
3k
500
1.2V
2N2222
9
8
10
7
1/2 LM319
SQUARE WAVE OUTPUT
1kHz to 1MHz
1N914
TRIANGLE WAVE OUTPUT
4
3
1
2
1/2 LM319
500pF
1k
1N914
3k
5V

FREQUENCY ADJUST MUST BE BUFFERED FOR R_L 10Ω

Fig. 97-18

FREQUENCY DIVIDER AND STAIRCASE GENERATOR



2N4381
C2
STAIRCASE OUTPUT
V_{IN}
C1
FJT1000
FJT1100
V+
7
2
3
6
4
V−
μA740
V+
9
1
8
7
10
6
5
μA734
V_{REF}
V−
V_{OUT}
0
V_{IN}
t
0
V_{REF}
7.0 V
0.2 V
nt

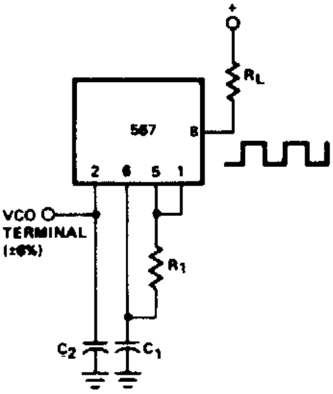
$$|V_{REF}| = 2V_D + N\left[3.5T + 2V_D - \frac{C_1 V_{IN}}{C_2}\right]$$

T in Seconds

V_D for FJT 1000 ≈ 0.31 V

Fig. 97-19

PRECISION OSCILLATOR TO SWITCH 100 mA LOADS



+
R_L
567
8
2
6
5
1
VCO TERMINAL (±6%)
R_1
C_2
C_1

Fig. 97-20

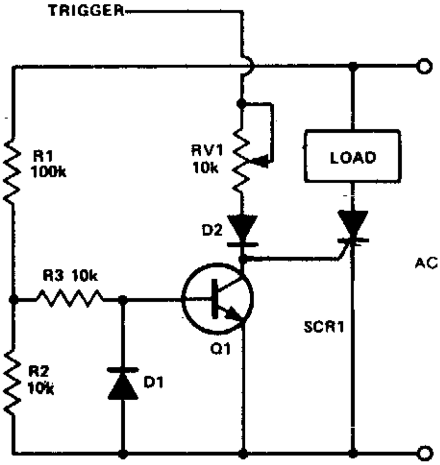
98

Zero Crossing Detectors

The sources of the following circuits are contained in the Sources section beginning on page 730. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

Zero Crossing Switch
Zero Crossing Detector
Zero Crossing Detector
Zero Crossing Detector with Temperature Sensor
Zero Crossing Detector
Zero Crossing Detector

ZERO CROSSING SWITCH


TRIGGER
R1
100k
RV1
10k
LOAD
D2
AC
R3 10k
SCR1
Q1
R2
10k
D1

Q1=GENERAL PURPOSE GERMANIUM
D1,2=GENERAL PURPOSE SILICON
SCR1=TO SUIT APPLICATION

Fig. 98-1

Circuit Notes

When switching loads with the aid of a thyristor, a large amount of RFI can be generated unless some form of zero crossing switch is used. The circuit shows a simple single transistor zero crossing switch. R1 and R2 act as a potential divider. The potential at their junction is about 10% of the ac voltage. This voltage level is fed, via R3, to the transistor's base. If the voltage at this point is above 0.2, the transistor will conduct, shunting any thyristor gate current to ground. When the line potential is less than about 2 V, it is possible to trigger the thyristor. The diode D1 is to remove any negative potential that might cause reverse breakdown.

ZERO CROSSING DETECTOR

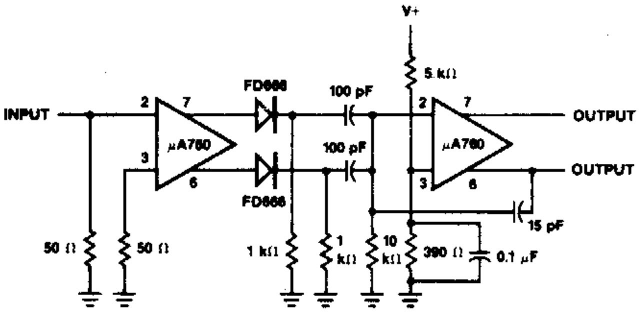
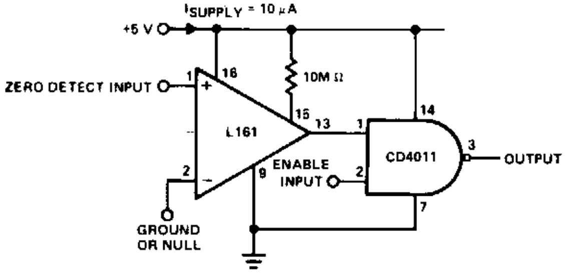

V+
5 kΩ
INPUT
2
7
μA760
3
6
FD666
100 pF
100 pF
FD666
2
7
μA760
3
6
OUTPUT
OUTPUT
15 pF
50 Ω
50 Ω
1 kΩ
1 kΩ
10 kΩ
390 Ω
0.1 μF

Fig. 98-2

Total Delay = 30 ns
Input frequency = 300 Hz to 3 MHz
Minimum input voltage = 20 mVpk-pk

ZERO CROSSING DETECTOR

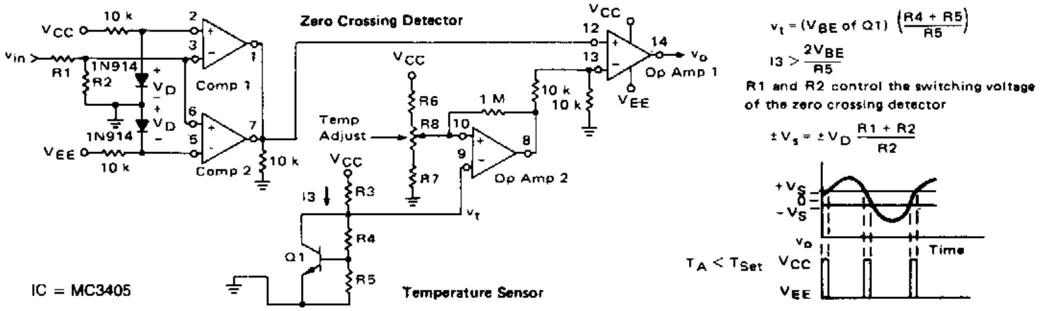

I_{SUPPLY} = 10 µA
+5 V
ZERO DETECT INPUT
1
16
10M Ω
15
13
14
L161
1
CD4011
3
OUTPUT
ENABLE INPUT
2
9
2
7
GROUND OR NULL

Circuit Notes

This detector is useful in sine wave squaring circuits and A/D converters. The positive input may either be grounded or connected to a nulling voltage which cancels input offsets and enables accuracy to within microvolts of ground. The CMOS output will switch to within a few millivolts of either rail for an input voltage change of less than 200 μV.

Fig. 98-3

ZERO CROSSING DETECTOR WITH TEMPERATURE SENSOR


V_{CC}
10 k
v_{in}
R1
1N914
R2
V_D
Comp 1
1N914
V_{EE}
10 k
Comp 2
10 k
Zero Crossing Detector
V_{CC}
R6
Temp Adjust
R8
R7
1 M
Op Amp 2
10 k
10 k
Op Amp 1
V_{CC}
V_{EE}
v_o
V_{CC}
R3
I3
Q1
R4
R5
v_t
Temperature Sensor
IC = MC3405
$+V_S$
0
$-V_S$
v_o
Time
$T_A < T_{Set}$
V_{CC}
V_{EE}

$$v_t = (V_{BE} \text{ of } Q1)\left(\frac{R4 + R5}{R5}\right)$$

$$I3 > \frac{2V_{BE}}{R5}$$

R1 and R2 control the switching voltage of the zero crossing detector

$$\pm V_s = \pm V_D \frac{R1 + R2}{R2}$$

Fig. 98-4

ZERO-CROSSING DETECTOR

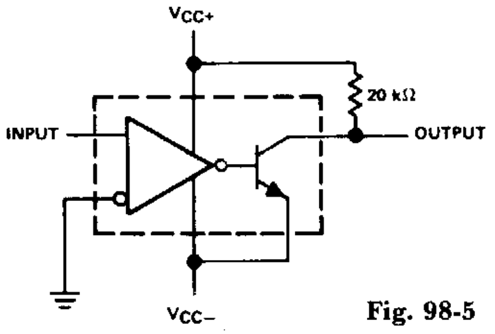
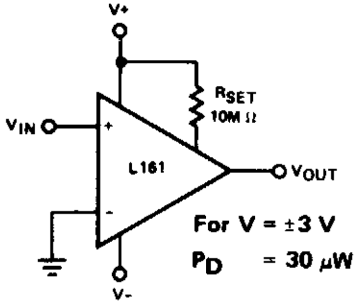

V_{CC+}
20 kΩ
INPUT
OUTPUT
V_{CC-}

Fig. 98-5

ZERO CROSSING DETECTOR


V+
R_{SET}
10M Ω
V_{IN}
L161
V_{OUT}
V-

For V = ±3 V

P_D = 30 µW

Fig. 98-6

Sources

Chapter 1

Fig. 1-1: The Build-It Book Of Electronics Projects, TAB Book No. 1498, p. 73.

Fig. 1-2: QST, 7/81, p. 28.

Fig. 1-3: Radio Electronics, 10/78, p. 41.

Fig. 1-4: '73 Magazine, 10/77, p. 122.

Fig. 1-5: Modern Electronics, 2/78, p. 50.

Fig. 1-6: Electronics Today International, 3/82, p. 69.

Fig. 1-7: Modern Electronics, 7/78, p. 51.

Fig. 1-8: Electronics Today International, 4/83, p. 72.

Fig. 1-9: 101 Electronic Projects, 1977, #64.

Fig. 1-10: Electronics Today International, 10/78, p. 94.

Fig. 1-11: Modern Electronics, 2/78, p. 55.

Fig. 1-12: Modern Electronics, 2/78, p. 48.

Fig. 1-13: Signetics 555 Timers, 1973, p. 26.

Fig. 1-14: Electronics Today International, 3/83, p. 23.

Fig. 1-15: Electronics Today International, 3/83, p. 23.

Fig. 1-16: National Semiconductor, Linear Databook, 1982, p. 3-288.

Fig. 1-17: Electronics Today International, 3/83, p. 23.

Fig. 1-18: Signetics 555 Timers, 1973, p. 22.

Fig. 1-19: 101 Electronic Projects, 1977, #65.

Fig. 1-20: Modern Electronics, 6/78, p. 58.

Fig. 1-21: Modern Electronics, 6/78, p. 55.

Chapter 2

Fig. 2-1: Modern Electronics, 3/78, p. 69.

Fig. 2-2: Electronics Today International, 10/78, p., 30.

Fig. 2-3: CQ, 5/77, p. 50.

Fig. 2-4: Ham Radio, 10/78, p. 34.

Fig. 2-5: Ham Radio, 10/78, p. 89.

Fig. 2-6: 73 Magazine, 7/78, p. 62.

Fig. 2-7: 101 Electronic Projects, 1975, p. 22.

Fig. 2-8: 73 Magazine, 7/82, p. 46.

Fig. 2-9: 73 Magazine, 7/83, p. 103.

Fig. 2-10: 101 Electronic Projects, 1975, p. 13.

Fig. 2-11: Ham Radio, 5/78, p. 87.

Fig. 2-12: 73 Magazine, p. 164.

Fig. 2-13: Modern Electronics, 2/78, p. 16.

Fig. 2-14: 73 Magazine, 10/77, p. 52.

Fig. 2-15: 73 Magazine, 7/77, p. 34.

Fig. 2-16: 104 Weekend Electronics Projects, TAB Book No. 1436, p. 120.

Fig. 2-17: Ham Radio, 10/70, p. 76.

Fig. 2-18: Electronics Today International, 7/77, p. 72.

Chapter 3

Fig. 3-1: Courtesy of Fairchild Camera & Instrument Corporation. Linear Databook, 1982, p. 4-119.

Fig. 3-2: Signetics Analog Data Manual, 1982, p. 3-83.

Fig. 3-3: Teledyne Semiconductor, Data & Design Manual, 1981, p. 11-207.

Fig. 3-4. Signetics Analog Data Manual, 1983, p. 10-99.

Fig. 3-5: Reprinted with the permission of National Semiconductor Corp. Data Conversion/Acquisition Databook, 1980, p. 3-107.

Fig. 3-6: Reprinted with the permission of National Semiconductor Corp. Transistor Databook, 1982, p. 11-29.

Fig. 3-7: Reprinted with the permission of National Semiconductor Corp. Audio/Radio Handbook, 1980, p. 2-67.

Fig. 3-8: Reprinted with the permission of National Semiconductor Corp. Hybrid Products Databook, 1982, p. 7-7.

Fig. 3-9: Electronics Today International, 2/82, p. 58.

Fig. 3-10. Signetics Analog Data Manual, 1983, p. 10-100.

Fig. 3-11: Precision Monolithics Incorporated 1981 Full Line Catalog, p. 12-50.

Fig. 3-12: Courtesy of Fairchild Camera & Instrument Corporation. Linear Databook, 1982, p. 9-17.

Fig. 3-13: Signetics Analog Data Manual, 1977, p. 35.

Fig. 3-14: Courtesy of Fairchild Camera & Instrument Corporation. Linear Databook, 1982, p. 5-39.

Fig. 3-15: Precision Monolithics Incorporated, 1981 Full Line Catalog, p. 6-10.

Fig. 3-16: Courtesy of Motorola Inc. Motorola Semiconductor Library, Volume 6, Series B, p. 8-21.

Fig. 3-17: Signetics Analog Data Manual, 1983, p. 17-17.

Fig. 3-18: Intersil Data Book, 5/83, p. 5-36.

Fig. 3-19: Courtesy of Motorola Inc. Linear Integrated Circuits, 1979, p. 3-17.

Fig. 3-20: Reprinted with the permission of National Semiconductor Corp. Hybrid Products Databook, 1982, p. 1-83.

Fig. 3-21: Precision Monolithics Incorporated, 1981 Full Line Catalog, p. 16-160.

Fig. 3-22: Signetics Analog Data Manual, 1982, p. 3-103.

Fig. 3-23: Precision Monolithics Incorporated, 1981 Full Line Catalog, p. 6-127.

Fig. 3-24: Courtesy of Motorola Inc., Linear Integrated Circuits, 1979, p. 3-83.

Fig. 3-25: Courtesy of Motorola Inc.

Linear Integrated Circuits, 1979, p. 3-131.

Fig. 3-26: Harris Semiconductor, Analog Data Book 1984.

Fig. 3-27: Intersil Data Book, 5/83, p. 5-36.

Fig. 3-28: Precision Monolithics Incorporated, 1981 Full Line Catalog, p. 16-37.

Fig. 3-29: Courtesy of Motorola Inc. Linear Integrated Circuits, 1979, p. 3-31.

Fig. 3-30: Siliconix Analog Switch & IC Product Data Book, 1/82, p. 6-21.

Fig. 3-31: Siliconix Analog Switch & IC Product Data Book, 1/82, p. 6-15.

Fig. 3-32: Precision Monolithics Incorporated, 1981 Full Line Catalog, p. 16-37.

Fig. 3-33: Siliconix Analog Switch & IC Product Data Book, 1/82, p. 7-56.

Fig. 3-34: Reprinted with permission of Analog Devices, Inc. Data Acquisition Databook, 1982, p. 4-119.

Fig. 3-35: Courtesy of Fairchild Camera & Instrument Corporation. Linear Databook, 1982, p. 4-42.

Fig. 3-36: Courtesy of Motorola Inc., Linear Integrated Circuits, p. 3-17.

Fig. 3-37: Courtesy of Motorola Inc. Linear Integrated Circuits, 1979, p. 6-23.

Fig. 3-38: Courtesy of Texas Instruments Incorporated. Linear Control Circuits Data Book, Second Edition, p. 145.

Fig. 3-39: Courtesy of Motorola Inc. Linear Integrated Circuits, 1979, p. 3-83.

Fig. 3-40: Courtesy of Fairchild Camera & Instrument Corporation. Linear Databook, 1982, p. 4-41.

Fig. 3-41: Canadian Projects Number 1, Spring/78, p. 29.

Fig. 3-42: Reprinted with the permission of National Semiconductor Corp. Application Note AN125, p. 2.

Fig. 3-43: Harris Semiconductor, Linear & Data Acquisition Products, p. 2-58.

Fig. 3-44: Reprinted with permission of Analog Devices, Inc. Data Acquisition Databook 1982, p. 4-98.

Fig. 3-45: Reprinted with the permission of National Semiconductor Corp. Application Note AN125, p. 3.

Chapter 4

Fig. 4-1: Courtesy of Fairchild Camera & Instrument Corporation. Linear Databook, 1982, p. 7-8.

Fig. 4-2: Intersil Data Book, 5/83, p. 4-83.

Fig. 4-3: Ferranti, Technical Handbook Vol. 10, Data Converters, 1983, p. 7-10.

Fig. 4-4: Precision Monolithics Incorporated, 1981 Full Line Catalog, p. 16-12.

Fig. 4-5: Reprinted with permission of Analog Devices, Inc. Data Acquisition Databook, 1982, p. 10-241.

Fig. 4-6: Precision Monolithics Incorporated, 1981 Full Line Catalog, p. 8-13.

Fig. 4-7: Reprinted with the permission of National Semiconductor Corp. National Semiconductor CMOS Databook. 1981. p. 3-63.

Fig. 4-8: Reprinted with permission of Analog Devices, Inc. Data Acquisition Databook, 1982, p. 10-240.

Fig. 4-9: Teledyne Semiconductor, Data & Design Manual, 1981, p. 7-39.

Fig. 4-10: Reprinted with permission of Analog Devices, Inc. Data Acquisition Databook, 1982, p. 10-50.

Fig. 4-11: Courtesy of Fairchild Camera & Instrument Corporation. Linear Databook, 1982, p. 5-32.

Fig. 4-12: Precision Monolithics Incorporated 1981 Full Line Catalog, p. 8-13.

Chapter 5

Fig. 5-1: Reprinted with the permission of National Semiconductor Corp. Data Conversion/Acquisition Databook, 1980, p. 3-22.

Fig. 5-2: Reprinted with the permission of National Semiconductor Corp. Transistor Databook, 1982, p. 11-29.

Fig. 5-3: Reprinted with the permission of National Semiconductor Corp. Data Conversion/Acquisition Databook, 1980, p. 8-64.

Fig. 5-4: Precision Monolithics Incorporated, 1981 Full Line Catalog, p. 12-39.

Chapter 6

Fig. 6-1: Electronics Today International, 3/82, p. 66.

Fig. 6-2: 101 Electronic Projects, 1977, IC 23.

Fig. 6-3: Reprinted with the permission of National Semiconductor Corp. Audio/Radio Handbook, 1980, p. 2-66.

Fig. 6-4: Electronics Today International, 10/79, p. 93.

Fig. 6-5: No reference.

Fig. 6-6: No reference.

Fig. 6-7: Electronics Today International, 3/75, p. 66.

Fig. 6-8: Electronics Today International, 3/78, p. 52.

Fig. 6-9: Electronics Today International, 5/78, p. 85.

Fig. 6-10: Modern Electronics, 7/78, p. 58.

Chapter 7

Fig. 7-1: Courtesy of Fairchild Camera & Instrument Corporation. Fairchild Semiconductor Application Note 300.

Fig. 7-2: Ham Radio, 1/78, p. 78.

Fig. 7-3: Courtesy of Motorola Inc. Linear Integrated Circuits, 1979, p. 6-23.

Fig. 7-4: 73 Magazine, 12/76, p. 97.

Fig. 7-5: 73 Magazine, 7/77, p. 34.

Fig. 7-6: Reprinted with the permission of National Semiconductor Corp. Linear Applications Handbook, 1982, p. AN29-9.

Fig. 7-7: Reprinted with the permission of National Semiconductor Corp. Linear Applications Handbook, 1982, P. LB16-1.

Fig. 7-8: Reprinted with the permission of National Semiconductor Corp. Transistor Databook, 1982, p. 11-31.

Fig. 7-9: Reprinted with the permission of National Semiconductor Corp. Linear Databook, 1982, p. 10-25.

Fig. 7-10: How to Design/Build Remote Control Devices TAB Book No. 1277, p. 230.

Fig. 7-11: Radio Electronics, 7/83, p. 7.

Fig. 7-12: Electronics Today International, Summer 1982, p. 45.

Fig. 7-13: 73 Magazine, p. 31.

Fig. 7-14: Reprinted from Electronics, 11/83. Copyright 1983, McGraw Hill Inc. All rights reserved.

Fig. 7-15: Electronics Today International, 7/72, p. 84.

Fig. 7-16: Courtesy of Motorola Inc. Linear Integrated Circuits, 1979, p. 3-42.

Fig. 7-17: Reprinted with the permission of National Semiconductor Corp. Linear Databook, 1982, p. 3-171.

Chapter 8

Fig. 8-1: Courtesy of Fairchild Camera & Instrument Corporation, Fairchild Progress, 11-12/76, p. 26.

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Fig. 97-18: Signetics Analog Data Manual, 1977, p. 264.

Fig. 97-19: Courtesy of Fairchild Camera & Instrument Corporation. Linear Databook, 1982, p. 5-25.

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Fig. 98-1: Electronics Today International, 8/78, p. 69.

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Fig. 98-3: ©Siliconix incorporated. Analog Switch & IC Product Data Book, 1/82, p. 6-18.

Fig. 98-4: Courtesy of Motorola Inc. Linear Integrated Circuits, 1979, p. 6-123.

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